BBC RD 1986/2

RESEARCH DEPARTMENT REPORT

Colour encoding and decoding techniques for line–locked sampled PAL and NTSC television signals

C.K.P. Clarke, B.Sc.(Eng), A.C.G.I.

Research Department, Engineering Division
THE BRITISH BROADCASTING CORPORATION

March,1986

BBC RD 1986/2

COLOUR ENCODING AND DECODING TECHNIQUES FOR LINE-LOCKED SAMPLED PAL AND NTSC TELEVISION SIGNALS

C.K.P. Clarke, B.Sc.(Eng.), A.C.G.I.

Summary

For many years subcarrier-locked sampling was preferred for digital processing of composite video signals. However, the use of line-locked sampling in composite signal coders and decoders is advantageous for compatibility with studio component signals, and for operation with monochrome signals and for colour signals in which the subcarrier to sync. frequency relationship is incorrect.

Digital techniques are described for generating, from a line-locked sampling frequency, quadrature subcarrier signals for use in composite colour encoders and for locking these signals to the incoming reference burst in colour decoders. This allows the advantages of the more stable, better defined performance obtained with digital implementation to extend to coders and decoders without the need for sample rate conversion to digital studio component signals. Although the main description is for the PAL system, modifications necessary for NTSC signals are also included.

Issued under the Authority of

M. E. B. Moffat

Head of Research Department

Research Department, Engineering Division
BRITISH BROADCASTING CORPORATION
March, 1986
(EL-181)

COLOUR ENCODING AND DECODING TECHNIQUES FOR LINE-LOCKED SAMPLED PAL AND NTSC TELEVISION SIGNALS

C.K.P. Clarke, B.Sc.(Eng.), A.C.G.I.

(EL-181)

COLOUR ENCODING AND DECODING TECHNIQUES FOR LINE-LOCKED SAMPLED PAL AND NTSC TELEVISION SIGNALS

C.K.P. Clarke, B.Sc.(Eng.), A.C.G.I.

1. Introduction

1.1. Digital signal coding in television studios

Digital signal coding, in which the television waveform is represented as a series of samples encoded as binary numbers, has considerable advantages for signal processing. In addition to protecting the signals against waveform distortion, binary coding allows digital arithmetic and storage techniques to be used. This greatly extends the range of processes which can be applied and ensures a stable and predictable performance.

Many of the parameters of digital signals for use in studios have been standardised in CCIR Recommendation 601.[1] This defines a form of coding based on line-locked sampling of the separate component luminance and colour difference signals Y, $R - Y$ and $B - Y$. An important result of locking the sampling to a harmonic of line frequency is that it produces an orthogonal grid of sample points on the picture. This means that filtering in each of the horizontal, vertical or temporal dimensions becomes a simple one-dimensional process. The use of separate component signals simplifies processing by avoiding the subcarrier phase variations which occur in composite encoded signals. Notwithstanding these advantages, it is only relatively recently that line-locked sampled component signals have been considered suitable as the basis for a studio signal standard.

In early digital television experiments, the PAL composite signals were sampled either at three-times the colour subcarrier frequency or at a nearby harmonic of line frequency.[2,3] Exactly which reference frequency was used depended on the application. If line-repetitive processing was involved such as in vertical interpolation,[4] line-locked sampling was favoured, while if similarly phased samples were required, such as in differential pulse code modulation (d.p.c.m.) prediction,[5] then subcarrier-locked sampling was used.

A refinement of subcarrier-locked sampling was to relate the sampling phases to the subcarrier reference phase,[6] in particular to specify that one sample should fall on the $+U$ axis. This had two advantages. First, it became possible to synchronise two digital colour signals for mixing by shifting the signals horizontally by one or two samples to match one another, and secondly, the phase of samples on adjacent lines was then related, in spite of the action of the PAL V-axis switch. For example, if samples OABOAB... were taken on one line (O being the 'on-axis' sample), then the samples would be in the order OBAOBA... on the adjacent lines. This feature simplified the synthesis of digital colour waveforms using a three-times subcarrier frequency clock because each colour could then be generated by repeating a sequence of three sample values.[7] Also, two dimensional prediction for d.p.c.m. was facilitated, because samples of similar subcarrier phase were available from the previous line.[8]

For a short period, three-times subcarrier frequency phase-locked sampling was widely used and regarded as a possible standard until a method of sampling PAL at the sub-Nyquist frequency of twice the colour subcarrier[9] was invented. In this method samples were taken at phases of 45° and 225° relative to the $+U$ axis. The attraction of the reduced data rate resulting from twice subcarrier sampling was judged to outweigh the small impairment involved, especially as this was found to be, in part, non-cumulative. Shortly afterwards another system of twice subcarrier frequency sampling was found[10] in which samples were taken at phases of 135° and 315°. This system produced digital Y and $(U + V/U - V)$ signals which could be converted back to PAL with almost no impairment.

The development of improved analogue-to-digital (a–d) and digital-to-analogue (d–a) converters introduced the possibility of sampling at four-times subcarrier frequency. By sampling with phases of 45°, 135°, 225° and 315° to the $+U$ axis, it was easy to produce both of the twice subcarrier systems of sampling. The digital comb filters used in the conversions from four-times subcarrier had better defined, more stable characteristics, so the resulting twice subcarrier sampled signals were improved in quality. The two systems of twice subcarrier sampling were thought to satisfy a wide range of requirements, including low data rate transmission[11] and reversible PAL coding and decoding,[12] with four-times subcarrier sampling providing a link between the two. In view of this, the two and four-times subcarrier sampling methods were suggested as the basis for a possible standard.[13]

However, further investigations revealed that the two systems of twice subcarrier sampling were not compatible.[14] When used in cascade, the two systems caused a serious loss of vertical chrominance and diagonal luminance resolution and largely negated the special features of the systems when used individually. Twice subcarrier sampling was therefore retained for digital PAL transmission, where the bit rate is at a premium, but in studio processing more emphasis was placed on four-times subcarrier frequency sampling. This sampling rate combines the advantages of subcarrier phase-related sampling with many of the features of line-locked sampling, notably the nearly orthogonal structure and the static grid of samples on the picture. Nevertheless, the extra storage capacity requirements and the reduced processing time in a clock period remained significant drawbacks.

1.2. PAL decoding with subcarrier-locked sampling

Throughout this period of development, a critical factor working in favour of subcarrier-locked sampling was the realisation that in some cases it would be too complicated to process the composite signal directly. Because of this, the signals would need to be decoded to component form. With samples taken at specific phases of the subcarrier, the colour content of each sample is known. For example, at four-times subcarrier with the samples locked to the 45° phases (preferred for the derivation of twice subcarrier samples), the four different phases contain $(U \pm V)/\sqrt{2}$, $(-U \pm V)/\sqrt{2}$, $(-U \mp V)/\sqrt{2}$ and $(U \mp V)/\sqrt{2}$, with the sign of the V component alternating with the PAL switch. At three-times subcarrier the weighting factors are some-what more complicated with the three phases containing U, $-\frac{1}{2}U \pm \sqrt{3}V/2$ and $-\frac{1}{2}U \mp \sqrt{3}V/2$. In either case, however, summing weighted proportions of the individual samples over a cycle of the subcarrier waveform can be used to extract both the colour difference signals and the luminance content. Thus digital PAL decoding with subcarrier-locked sampling was considered to be a relatively simple matrixing operation. This ignored the fact that a fundamental element in the 'digital' decoder was, in fact, the analogue phase-locked sampling process.

Subcarrier-locked decoding, nevertheless, was not without its drawbacks. It perpetuated the complications of digital PAL processing by carrying the subcarrier relationship into the YUV component samples. Twice and three-times subcarrier frequency sampling produced moving sample grids which were much less convenient for component signal processing than the static orthogonal samples of a line-locked system. With the alternative of four-times subcarrier sampling, although the sampling grid was static and nearly orthogonal, there was an uncertainty in its position relative to the picture information, since this depended on the undefined subcarrier-to-line phase relationship at the original PAL coder.

Further problems were caused if non-mathematical PAL signals were encountered, since these signals do not have the correct subcarrier frequency to line frequency relationship. In this case, because the sampling structure follows the subcarrier, the picture information drifts sideways until reset with a jerk when sufficiently out of step with the line pulses. Although the horizontal movement amounts to only one sample width, the repeated drift and jerk is very disturbing. Also, with monochrome signals, in which the colour burst is omitted to prevent cross-colour on colour receivers, there is no subcarrier available as a reference for the sampling loop.

The absence of a satisfactory solution to the problems of subcarrier-locked decoding finally weighed against having a studio standard based on digitised PAL signals. Also it was recognised that so many processes would benefit from using digital component signals, preferably in line-locked form, that it was better that the signals should remain in component form rather than being converted to and from PAL. The task of PAL decoding was therefore relegated to one of interfacing the existing analogue PAL system to the new line-locked digital component signals during a changeover period.

1.3. PAL decoding with line-locked sampling

With line-locked sampling there is no simple relationship between individual samples and the subcarrier reference phase, so the simple matrixing method for subcarrier-locked sampling cannot be applied. Indeed, because of this it was at one time widely held that digital PAL decoding was impracticable unless subcarrier-locked sampling was used.[13,15] Digitising the outputs of an analogue PAL decoder with three a–d converters using line-locked sampling seemed the only economically viable method of obtaining line-locked component signals from PAL, but this necessarily lost the performance benefits of digital processing techniques in the decoder. However, the problems of digital decoding in a line-locked sampling system were confronted during the development of an experimental converter working between the 625/50 PAL and 525/60 NTSC television standards.[16]

The converter itself[17] used two closely related line-locked sampling frequencies for the input and

output standards. The frequencies were chosen to make the number of active-line samples the same on the two standards, so avoiding the need for any horizontal interpolation in the converter.

The use of analogue decoding during these experiments was not favoured because several quite complicated comb filtering methods were to be compared and analogue parameter variations could have affected the results. The possibility of encountering non-mathematical composite signals also ruled out the use of subcarrier-locked sampling, even if two appropriately related frequencies had been available. The convenience of operating digital decoders and coders directly at the line-locked sampling frequencies of the converter was therefore very attractive.

The method of decoding developed for this application differed from the prevailing understanding of digital decoding in that the analogue sampling process made no contribution to demodulating the colour signals. Instead all the processes of generating a local reference subcarrier, locking it to the incoming burst phase with a feedback loop and extracting the colour signals using product demodulators were carried out by digital circuitry.[18,19] It was found that similar techniques could be used for PAL and NTSC, and could be applied to composite encoders as well. Indeed, the confirmation that these line-locked techniques provided a viable and cost-effective solution for coding and decoding proved to be a significant step in support of standardisation on the use of line-locked sampled digital component signals for studios.[20] Since then, similar techniques have been used in more general investigations aimed at optimising the luminance-chrominance separation filters for digital studio inputs. The development of these separation filters will be described in a subsequent Report.

The technique used to generate subcarrier signals from a line-locked sampling frequency is fundamental both to digital coders and decoders. In the following description, therefore, this is treated first, followed by the explanation of modulation and its application in coders. In decoders, it is advantageous to apply digital techniques to control the sampling loop for analogue-to-digital conversion of the composite signal. The process of demodulation, although similar to modulation in many respects, has the additional requirement to lock the local reference subcarrier to the incoming burst phase. The description of this process includes an explanation of the way in which a line-locked sampled demodulator is able to accommodate non-mathematical PAL signals. Finally, the application of special techniques such as PAL modifiers in decoders is described. Throughout the text, although the main description deals with the PAL system, the differences necessary for NTSC operation are also included.

2. Subcarrier generation from a line-locked sampling frequency

Colour coders and decoders both use quadrature subcarrier signals locked to the subcarrier reference phase. In a digital circuit, a train of samples representing a subcarrier waveform is generated by selecting an appropriate sinusoidal sample value corresponding to the current phase of the reference. If subcarrier phase-locked sampling is used, then each sample is associated with a defined value of the reference phase. This can be monitored by a simple counter, for example, with three-times subcarrier sampling, the phases 0°, 120° and 240° could be identified by counter states 0, 1 and 2. Samples values corresponding to 0, $\frac{\sqrt{3}}{2}$ and $-\frac{\sqrt{3}}{2}$ for a sine wave or 1, $-\frac{1}{2}$ and $-\frac{1}{2}$ for a cosine wave could then be selected, as shown in Fig. 1.

With line-locked sampling there is no direct relationship defined between individual samples and

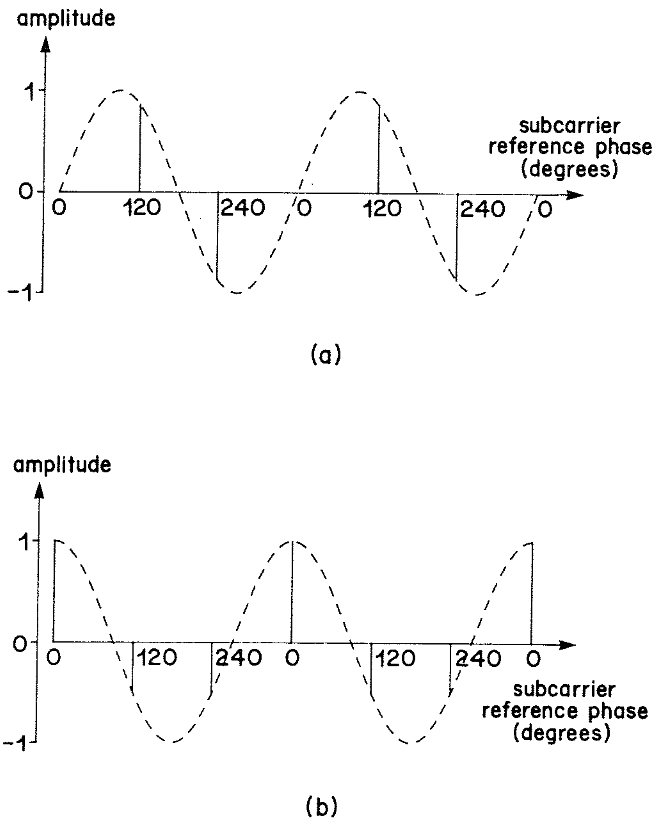


Fig. 1 – Three-times subcarrier phase-locked sampling: samples of subcarrier frequency waveforms generated from the subcarrier reference phase producing (a) a sine wave and (b) a cosine wave.

the subcarrier reference phase. However, the subcarrier and sampling frequencies are still related, although in a complicated way.

2.1. Subcarrier phase generation

2.1.1. Frequency relationships

In the PAL (System I) and NTSC (System M) colour standards there are prescribed relationships[21] between the subcarrier frequency (f_{sc}) and the line frequency (f_h). For PAL:

$$\frac{f_{sc}}{f_h} = \frac{709379}{2500}$$

Also, the 13.5 MHz sampling rate (f_s) of the digital studio standard is related to the line frequency,[1] as shown:

$$f_s = 864 f_h$$

Combining these equations produces the following relationship between the subcarrier frequency and the line-locked sampling frequency of the digital studio standard:

$$\frac{f_{sc}}{f_s} = \frac{709379}{2160000}$$

which can also be expressed in terms of the sampling period (t_s) and the subcarrier period (t_{sc}):

$$\frac{t_s}{t_{sc}} = \frac{709379}{2160000}$$

In each clock period, therefore, the subcarrier phase advances by this fraction of a subcarrier cycle.

The following somewhat simpler relationships are obtained for NTSC:

$$\frac{f_{sc}}{f_h} = \frac{455}{2}$$

and[1]

$$f_s = 858 f_h$$

so that

$$\frac{f_{sc}}{f_s} = \frac{35}{132}$$

or

$$\frac{t_s}{t_{sc}} = \frac{35}{132}$$

Similar relationships could be obtained for any other line-locked sampling frequencies, although in most cases these would not provide a common sampling frequency (13.5 MHz) on the two standards.

2.1.2. Ratio counters

It is possible to design a counting circuit[7,22] based on any ratio in the form $p:q$. This consists of an accumulator in which the smaller number p is added at each clock pulse modulo another number q. In hardware, the counter consists of an adder and a register as shown in Fig. 2. The contents of the register are constrained so that if they would exceed or equal q, q is subtracted from the contents.

In the present application, this overflow corresponds to the completion of a full cycle of subcarrier. Since only the remainder, that is, the subcarrier phase, is required, the number of whole cycles completed is of no interest. During each clock period, the output of the register shows the relative phase of a subcarrier frequency waveform in $q$ths of a subcarrier period. So, by using the register contents to address a read-only memory (ROM) containing a sine wave characteristic, a numerical representation of the sampled subcarrier sine wave can be produced.

While the 132 phases required for NTSC are perfectly manageable, the PAL ratio would require a ROM with 2,160,000 words capacity. This can be avoided by partitioning the ratio into two fractions, the more significant of which provides the subcarrier reference phase. In these circumstances it is advantageous to make the denominator of the more significant fraction a power of two, partly because the overflow of the ratio counter is then automatic and partly to use the full capacity of the ROM.

If the subcarrier period is divided into 2048 phase steps, then the partitioned ratio for PAL becomes:

$$\frac{t_s}{t_{sc}} = \frac{672\frac{10064}{16875}}{2048}$$

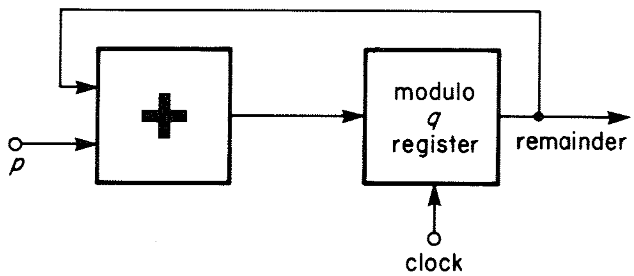


Fig. 2 – A p:q ratio counter.

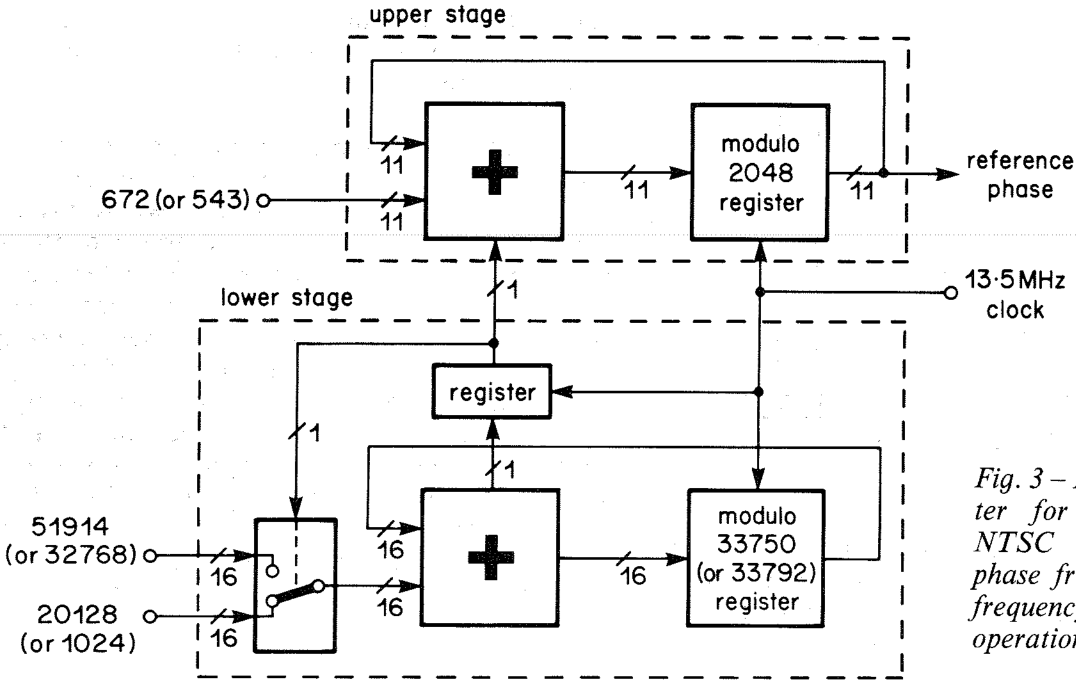


Fig. 3 – A two-stage ratio counter for generating PAL or NTSC subcarrier reference phase from a 13·5 MHz clock frequency. Values for NTSC operation are shown in parentheses.

or, if a 16-bit accumulator is used for the less significant ratio:

$$\frac{t_s}{t_{sc}} = \frac{672\frac{20128}{33750}}{2048}$$

The partitioning can also be applied to NTSC to ensure automatic overflow in the counter feeding the ROM and to achieve greater commonality of processing for the two systems. Then:

$$\frac{t_s}{t_{sc}} = \frac{543\frac{1}{33}}{2048}$$

which, with a 16-bit accumulator, becomes:

$$\frac{t_s}{t_{sc}} = \frac{543\frac{1024}{33792}}{2048}$$

In this form, the corresponding ratio counter has two stages as shown in Fig. 3. The less significant stage produces a sequence of carry bits which correct the approximate ratio of the upper stage by altering the counting step from 672 to 673 (543 to 544 for NTSC). The upper stage then produces an accurate 11-bit reference phase output to address a ROM.

While the upper stage adder automatically overflows to provide modulo 2048 operation, the lower stage requires additional circuitry because 33750 (33792 for NTSC) is not an integer power of two. In this case, the 16-bit register has a maximum capacity of 65535 and the adder generates a carry for any value greater than this. To produce the correct carry sequence, it is necessary, each time the adder overflows, to adjust the next number added to make up the difference between 65536 and 33750 (33792). This requires:

$$65536 - 33750 + 20128 = 51914 \text{ for PAL}$$

or $$65536 - 33792 + 1024 = 32768 \text{ for NTSC}$$

Although this changes the contents of the lower stage register, the sequence of carry bits is unchanged, so ensuring that the correct phase values are generated.

2.2. Quadrature subcarrier generation

Each value of the 11-bit reference phase signal corresponds to one of 2048 waveform values taken at a particular point in the subcarrier cycle period and stored in a ROM. The maximum phase error produced by this quantisation is $\pm 0.09°$, which causes a maximum amplitude error of $\pm 0.075\%$ (relative to the peak-to-peak amplitude) at the steepest part of the sine wave signal. In comparison, quantisation of a standard level composite signal to 8 bits in an a–d converter causes $\pm 0.28\%$ error in the largest possible peak-to-peak amplitude of subcarrier. Therefore, in a demodulator, quantisation of the reference phase signal is a minor contribution to the total quantisation distortion. Even so, a further reduction of distortion could be obtained by using a 12-bit phase signal, although twice the ROM capacity would then be required to store the subcarrier wave-shape.

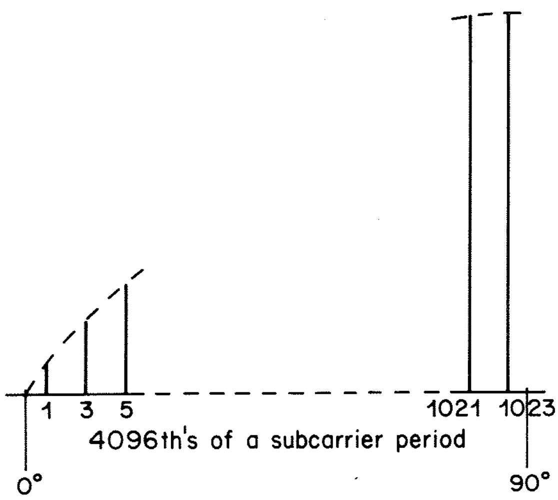


Fig. 4 – Positions of stored sample values for one quadrant of a subcarrier cycle. Samples for other quadrants can be generated by inverting the addresses and/or the sample values.

The ROM requirements can be minimised by storing only one quadrant of the subcarrier waveshape for each of the quadrature signals. The values for the other quadrants can then be produced using the symmetrical properties of the sinusoidal waveshape. This requires the 512 values used to store a quadrant to be positioned symmetrically throughout a 90° section of the characteristic. Thus, rather than taking the sample values at the 2048ths points on the waveform, they are taken at odd multiples of one 4096th of the total period, as shown in Fig. 4. This avoids end-effects when the sample values are read out in the reverse order.

Fig. 5 shows a circuit arrangement for generating quadrature subcarriers from an 11-bit subcarrier reference phase signal. This uses two ROM units with 9-bit addresses to store quadrants of the sine and cosine waveforms. Exclusive-OR gates are used to invert the addresses for generating time-reversed portions of the waveforms and to invert the output polarity to make negative portions of the waveforms. An additional gate is included in the sign bit for the V subcarrier to allow injection of a PAL square wave signal to provide phase inversion of the V signal on alternate lines. The output word length of the ROM's used to generate the subcarrier samples tends to be limited more by the following circuitry than by constraints imposed by the ROM size.

A further feature that is often conveniently included in the ROM is a gain change. By storing a scaled version of the waveform, the gain factors converting from, for example, $B - Y$ to U or $R - Y$ to V can be accommodated. This is discussed further in Sections 3.2.2 and 4.3.3.

3. Colour encoding

In addition to the quadrature subcarrier generation described in the previous Section, the other

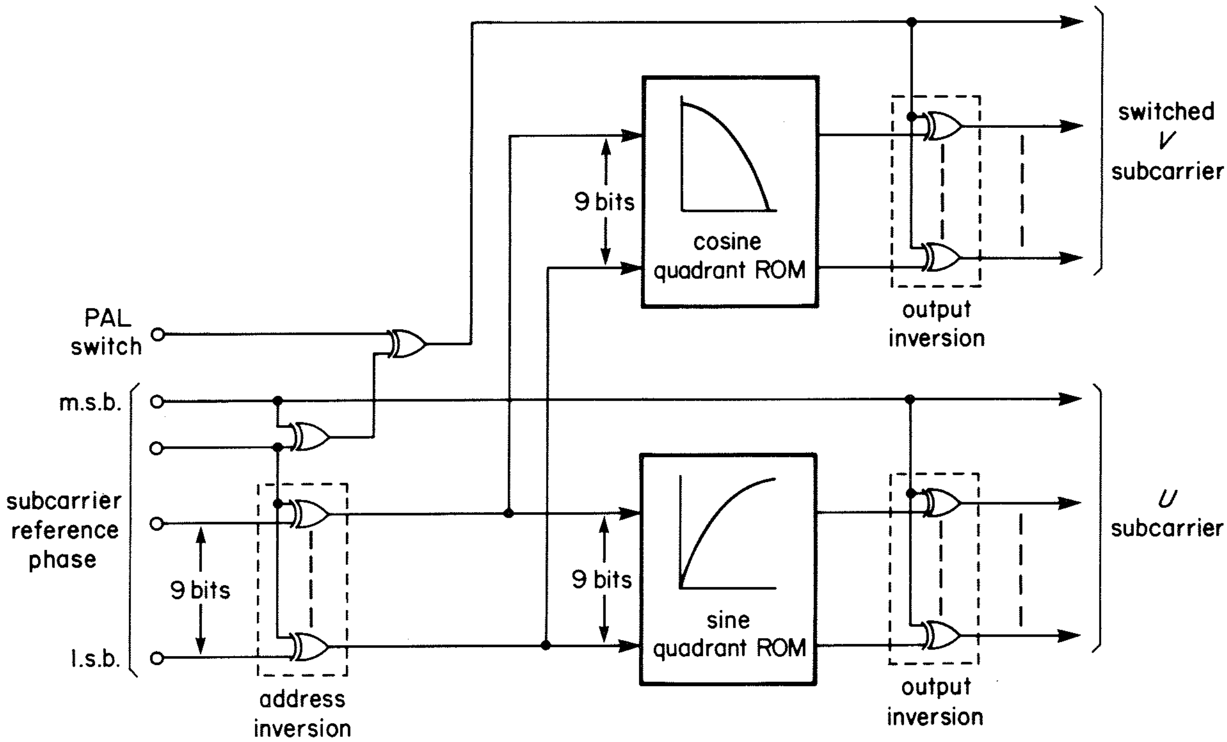


Fig. 5 – Quadrature subcarrier generation.

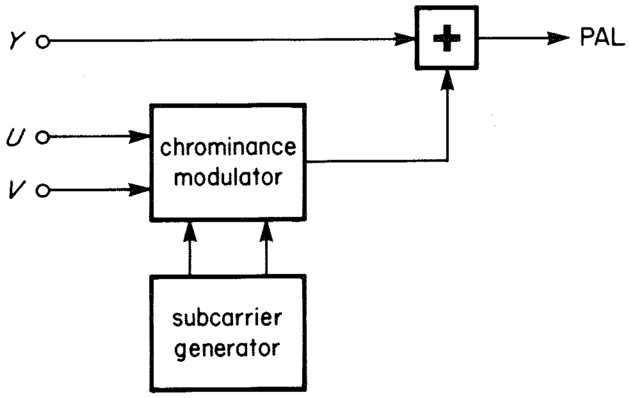


Fig. 6 – Basic processes of a PAL coder.

basic process of a conventional PAL coder is chrominance modulation. The weighted colour difference signals are modulated onto the high frequency subcarrier signals and added to the luminance as shown in Fig. 6. Although no compensating delay is shown, the luminance and modulated chrominance signals should be co-timed at the point of addition. The same basic arrangement is used for NTSC although there are minor differences as described later in this Section. Also mentioned later are 'clean' coding methods for PAL and NTSC which involve additional filtering in the coder.

3.1. Features of the PAL signal

The PAL system uses double sideband suppressed-carrier amplitude modulation of two subcarriers in phase quadrature to carry the two weighted colour difference signals, U and V. Because the subcarriers are orthogonal, the U and V signals can be separated perfectly from each other. However, if the modulated chrominance signal is distorted, either through asymmetrical attenuation of the sidebands or by differential phase distortion, the orthogonality is degraded, resulting in crosstalk between the U and V signals. As this is quite likely to occur, the PAL system incorporates alternate line switching of the V signal component. This is an additional form of modulation which provides a frequency offset between the U and V subcarriers in addition to the phase offset. Thus, when decoded, any crosstalk components appear modulated onto the alternate line carrier frequency, in plain coloured areas producing the moving pattern known as Hanover bars. This pattern can be suppressed in the decoder by a comb filter averaging equal contributions from switched and unswitched lines.

The PAL chrominance signal can be represented mathematically as a function of time (t) by the

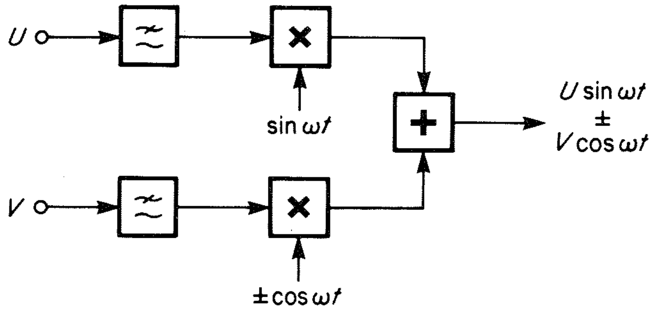


Fig. 7 – PAL chrominance modulation.

expression:

$$U \sin \omega t \pm V \cos \omega t$$

where $\omega = 2\pi f_{sc}$

and for System I PAL signals

$$f_{sc} = 4.43361875\,\text{MHz}$$

The sign of the V component alternates from one line to the next. This signal is formed by the product modulators of Fig. 7 which multiply the orthogonal subcarrier waveforms by the instantaneous values of low-pass filtered baseband U and V signals. The low-pass filters use a frequency characteristic approximating to the Gaussian template shown in Fig. 8.

The modulation process is shown in spectral terms in Fig. 9. Fig. 9(a) represents the baseband spectrum of a full bandwidth colour difference signal, the high frequency components of which are attenuated by low-pass filtering to produce the spectrum of Fig. 9(b). When convolved with the line spectrum of the subcarrier signal, Fig. 9(c), this produces the modulated chrominance signal spectrum of Fig. 9(d).

A Gaussian frequency characteristic (Fig. 8) is used for the colour difference low-pass filters for optimum compatibility with monochrome receivers. This wide-band, slow roll-off characteristic is needed to minimise the disturbance visible at the edges of coloured objects because, with a monochrome display, the signal receives no further filtering. A narrower, sharper-cut characteristic would emphasise the subcarrier signal at these edges, widening the transitions and introducing ringing.

Because the coder uses wide-band U and V filters, the modulated chrominance spectra can overlap to cause aliasing, as shown in the zero frequency region of Fig. 9(d). Also, if the PAL coder

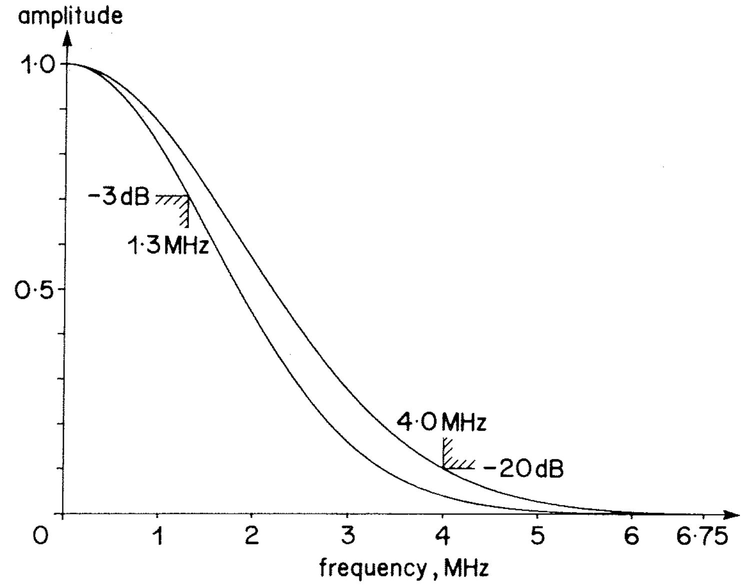


Fig. 8 – Gaussian low-pass filter characteristics for use in PAL coders.

does not contain a 5.5 MHz low-pass filter at the output, the upper chrominance sidebands extend well beyond normal video frequencies. In practice, the PAL signal may sometimes retain these out-of-band components up to the broadcast transmitter, at which point the portion of the upper sideband above 5.5 MHz is removed.

Although the coder maintains a wide chrominance bandwidth, the colour difference signal bandwidth reproduced in the decoder is usually much narrower than this. This is because if the decoder reproduces colour signal frequencies above 1.07 MHz, the loss of the upper sidebands above 5.5 MHz leads to ringing and *U-V* crosstalk on colour transitions. These effects are caused, respectively, by the sharp-cut characteristic of the 5.5 MHz filter and the sideband asymmetry. Also, any increase in the decoder bandwidth causes a proportionate increase in cross-colour, whilst the additional chrominance resolution obtained diminishes rapidly as the bandwidth is increased.

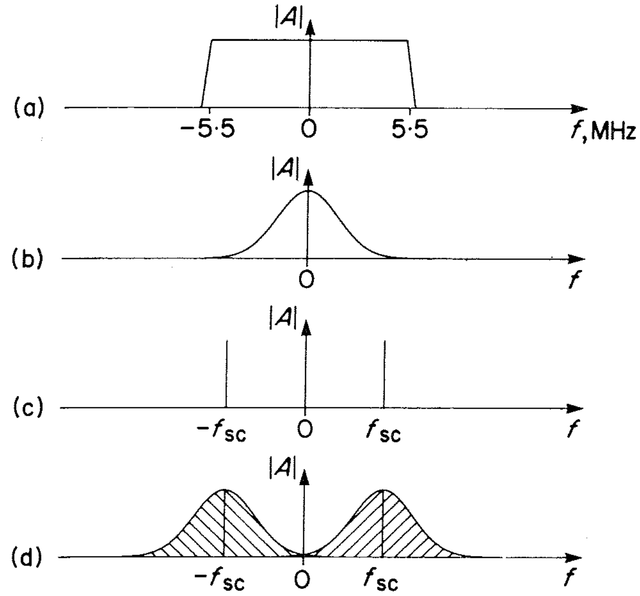


Fig. 9 – Frequency spectra in PAL chrominance modulation: (a) the baseband colour difference signal, (b) the Gaussian filtered colour difference signal, (c) the subcarrier sinewave and (d) the modulated chrominance spectrum produced by convolving (b) and (c).

In view of these considerations, if compatibility with monochrome displays could be ignored, the encoded chrominance bandwidth could be reduced without losing any resolution for the colour viewer. On the contrary, it would have a beneficial effect for the colour viewer because the spread of the lower chrominance sidebands into the low frequency luminance region would be reduced. This, in turn, would limit the amount of low frequency cross-luminance, which is difficult to suppress in a decoder.

3.2. Digital PAL encoding

Whereas an analogue PAL coder depends on several waveforms from an associated synchronising pulse generator, it is usually convenient to incorporate such a generator as part of a digital coder. Two alternative arrangements are shown in Fig. 10. Fig. 10(a), for analogue *RGB* input signals, contains a–d converters and a 13.5 MHz clock pulse generator locked to the line frequency of the incoming synch-

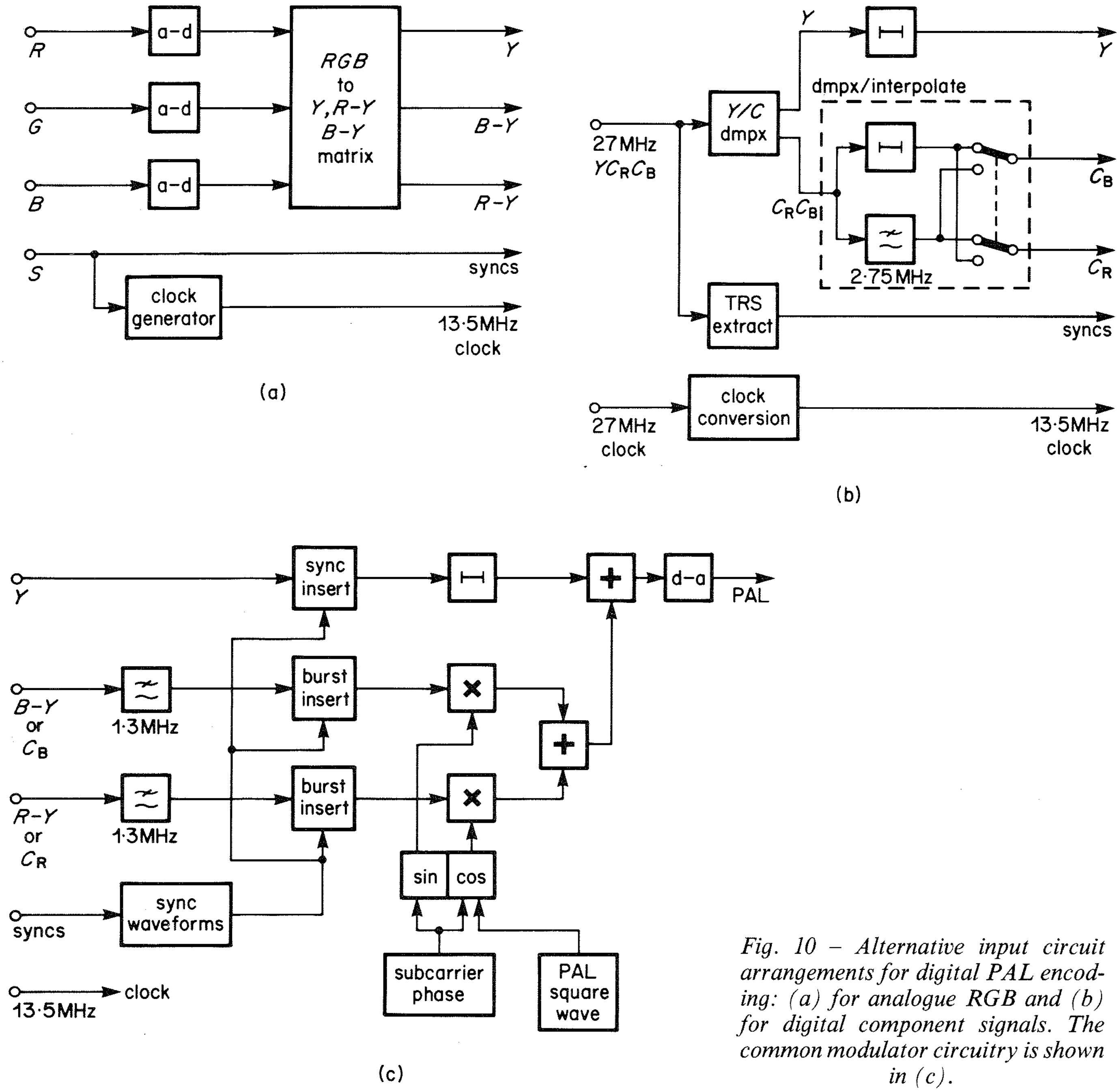


Fig. 10 – Alternative input circuit arrangements for digital PAL encoding: (a) for analogue RGB and (b) for digital component signals. The common modulator circuitry is shown in (c).

ronising pulses. The digitised *RGB* signals are then matrixed to form luminance and colour difference components. Alternatively, Fig. 10(b) shows the demultiplexing and sample interpolation circuitry[23] needed to convert from a 27 MHz multiplexed digital component signal. In this case, synchronising information is obtained from the timing reference signal (TRS) codes[24] and 13.5 MHz clocks are produced from division of the 27 MHz clock carried with the digital component signal. The remaining circuitry shown in Fig. 10(c), including sync. and burst insertion, the digital transversal low-pass filters and the four-quadrant digital multipliers, is common to both arrangements.

Although in most respects the alternatives of Fig. 10 would produce substantially similar results, the *RGB* input arrangement of Fig. 10(a) has a significant advantage. The PAL System I specification requires that the composite signal should represent only signals within the normal ranges of *R*, *G* and *B*.[25] The *RGB* inputs effectively ensure this because any signals extending significantly beyond the normal range will be clipped by the a–d converters. Alternatively, digital *RGB* inputs could be clipped in a similar manner. The digital YC_RC_B input, however, does not have this facility, since the only way of finding whether the YC_RC_B combinations will produce valid *RGB* values is to convert to

RGB. The responsibility for ensuring that the combinations are valid is therefore passed to the YC_RC_B system as a whole. In practice, this may be difficult to control unless processes such as graphics, which could produce unreal colours when operating in YC_RC_B, always operate with *RGB* signals.

3.2.1. Modulation

When a coder is implemented with digital circuitry, each signal is represented by a stream of binary coded samples. Because of this, the frequency spectra of the modulation process are those of the analogue coder (Fig. 9) repeated at harmonics of the sample rate as shown in Fig. 11. With a sample rate of 13.5 MHz, the upper sidebands of the modulated signal extend beyond the Nyquist limit and can cause high frequency chrominance to be aliased to fall near to the subcarrier frequency. For a digital coder, then, the use of a rather narrower pre-modulation low-pass filter would be more suitable to reduce the alias caused by 13.5 MHz sampling.

When the 13.5 MHz modulating signals are obtained from a digital 4:2:2 YC_RC_B input as shown in Fig. 10(b), the colour difference signals will have already been attenuated above 2.75 MHz, with alias components from the 6.75 MHz sampling frequency present between 2.75 and 4 MHz. The spectrum of these signals is shown in Fig. 12(a). The aliased region is then further attenuated by the Gaussian filters of the modulator. However, when modulated, this region is aliased again by the 13.5 MHz sampling frequency to produce components between 5.07 and 6.32 MHz. This is shown in Fig. 12(b) with the effect of the Gaussian filters omitted for clarity. As many of these components are above 5.5 MHz, the characteristic of the d–a converter low-pass filter shown in Fig. 12(c) will provide some additional attenuation at the coder output. Overall, therefore, the effect of the alias components due to 6.75 MHz sampling is negligible. However, the loss of the upper sidebands above 5.5 MHz still results in ringing and *U-V* crosstalk as described in Section 3.1.

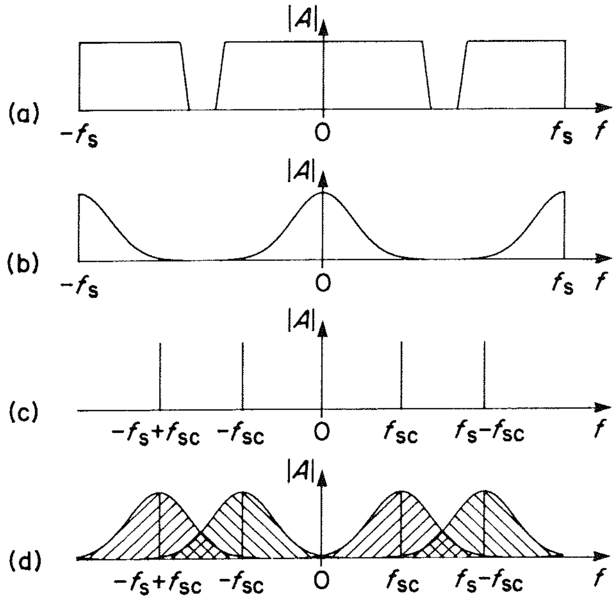


Fig. 11 – Frequency spectra in digital PAL chrominance modulation: (a) the baseband colour difference signal, (b) the Gaussian filtered colour difference signal, (c) the subcarrier sinewave and (d) the modulated chrominance spectrum produced by convolving (b) and (c).

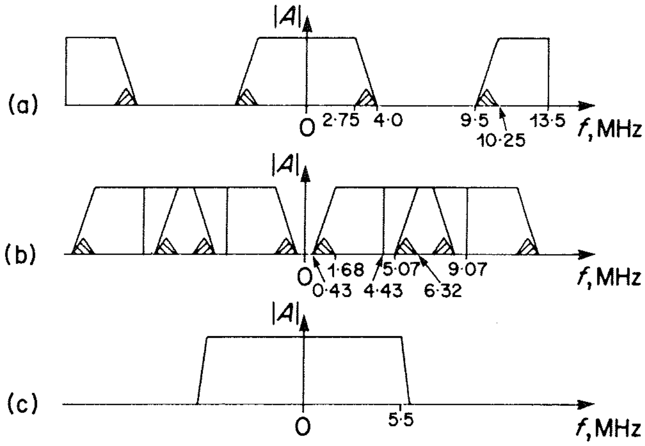


Fig. 12 – Frequency spectra for digital PAL chrominance modulation with signals derived from C_R and C_B signals sampled at 6.75 MHz: (a) the spectrum of a 13.5 MHz sampled colour difference signal derived from 6.75 MHz samples showing the region of attenuation and aliasing (shaded); (b) the positions of the spectral components after modulation, ignoring the effect of the coder Gaussian filters; and (c) the video passband region producing additional attenuation above 5.5 MHz.

3.2.2. Signal relationships in coders

The PAL encoding process involves a number of signal conversion and scaling operations to produce the composite signal from either *RGB* or digital YC_RC_B component signals. If converting from analogue *RGB*, it is convenient to make the coding range at the a–d converters 16 to 235 for a standard level signal, to accord with CCIR Recommendation 601.[1] The *Y* signal can be obtained from *RGB* using the following relationship:[21]

$$Y = 0.299R + 0.587G + 0.114B$$

However, *Y* can be calculated more conveniently

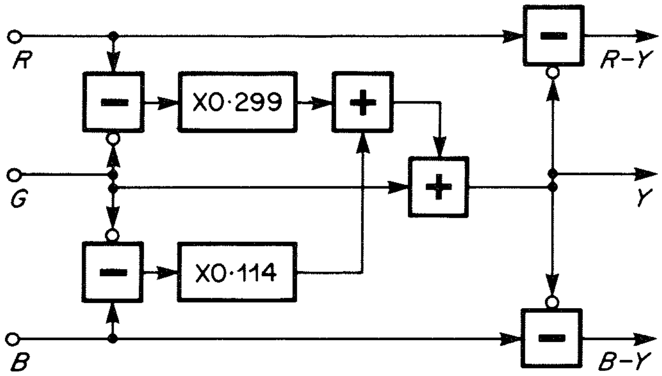


Fig. 13 – A simplified matrix arrangement for deriving Y, R – Y and B – Y signals from R, G and B.

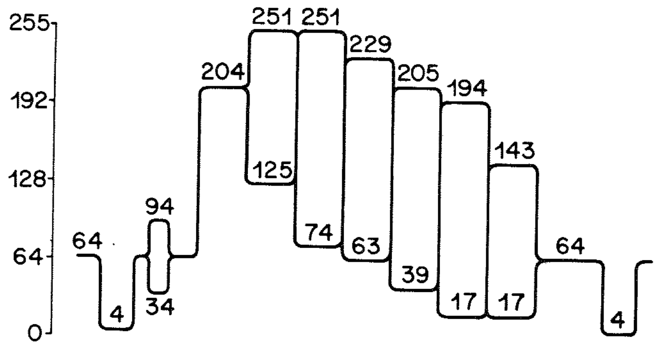


Fig. 14 – Eight-bit coding levels in a 100% colour bars composite PAL signal.

from the rearranged form:[26]

$$Y = 0.299(R - G) + G + 0.114(B - G)$$

with the colour difference signals $R - Y$ and $B - Y$ being obtained subsequently by subtraction. The matrix arrangement based on this relationship is shown in Fig. 13.

Alternatively, $R - Y$ and $B - Y$ can be obtained as shown in Fig. 10(b) from the digital colour difference signals C_R and C_B by the following relationships:[1]

$$B - Y = \frac{0.886}{0.5} \cdot \frac{219}{224} \cdot C_B$$

$$R - Y = \frac{0.701}{0.5} \cdot \frac{219}{224} \cdot C_R$$

Further gain changes are required to convert the $R - Y$ and $B - Y$ signals to the weighted colour difference signals U and V:[21]

$$U = 0.493(B - Y)$$

and $$V = 0.877(R - Y)$$

and an additional factor of 0.639, that is 140/219, must be included in Y, U and V to optimise the coding range of the composite PAL signal for d–a conversion.

For the colour difference signals it is advantageous to combine these two sets of scaling factors into a single factor for each signal. These can be applied conveniently by weighting the subcarrier sine and cosine values stored in each ROM, although the colour burst amplitudes have to be adjusted to take account of the subsequent gain factor. The luminance path, however, requires a dedicated ROM for gain conversion and to alter the black level code from 16 to 64 in the composite output signal. The output levels for a composite PAL 100% colour bars signal are then as shown in Fig. 14 with sync. bottom at level 4 and chrominance at the top of the yellow bar reaching level 251. Although Fig. 14 shows codes appropriate for an 8-bit signal, it is advantageous to carry additional lower significance bits through from the modulators. Then a 10-bit d–a converter can be used with only a small increase in cost.

3.2.3. Synchronising pulse and colour burst generation

Although the generation of subcarrier waveforms is more complicated with line-locked sampling than it would be with subcarrier-locked sampling, the generation of substantially line-repetitive waveforms such as synchronising pulses, blanking edges and the colour burst envelope is greatly simplified. These waveforms are generated from the 13.5 MHz clock frequency by a combination of sample-rate and line-rate counters, for PAL, dividing by 864 and 625 respectively. The two counters, shown in Fig. 15(a), are locked to line and picture pulses derived from either mixed syncs, in the case of an analogue *RGBS* signal, or the TRS code, for a digital component input. In the absence of an input signal, the counters can free-wheel so that standard sync. and burst waveforms are maintained at the output.

The method used to generate the mixed sync., burst gate and mixed blanking waveforms was originally developed* for use in an electronic zone plate generator based on line-locked sampling.[27] The two-level control waveforms are produced by the system of read-only memories shown in Fig. 15(b) driven by the sample and line-rate counters.

* The method was subsequently enhanced by P.W. Fraser.

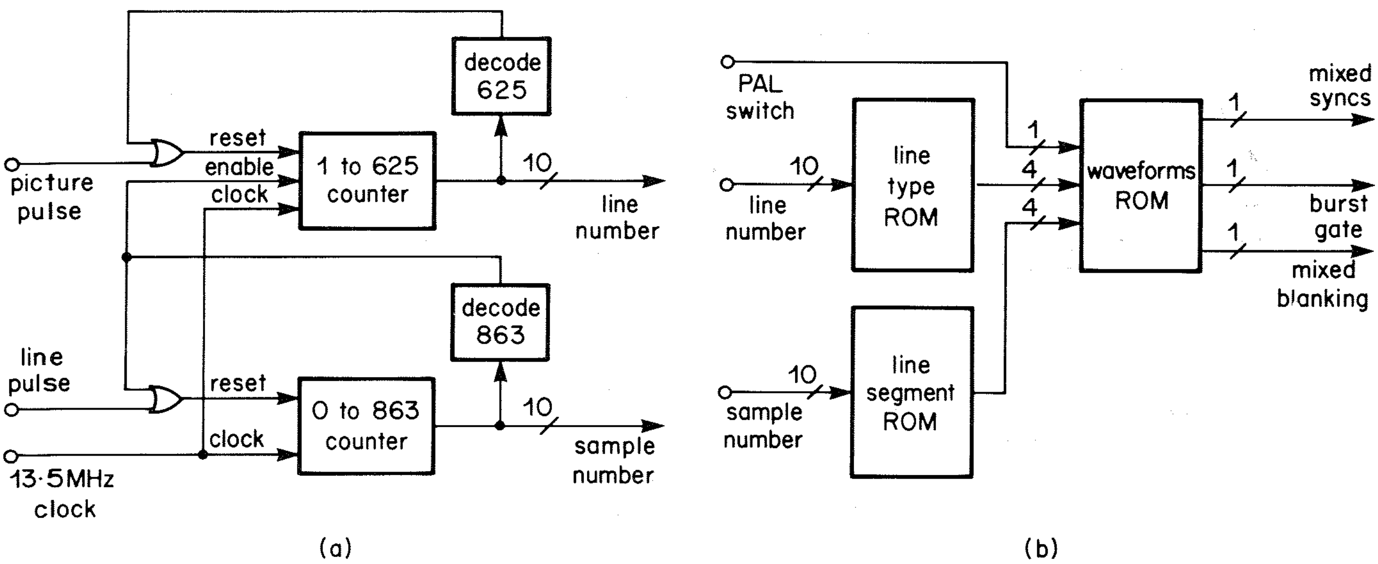


Fig. 15 – Line-repetitive waveform generation for the 625/50 system: (a) sample and line-rate counters and (b) the arrangement of rom's used to generate sync., burst and blanking waveforms.

One ROM converts the sample numbers into codes representing the periods between all possible transition points of the three waveforms over the line period. These are shown in Fig. 16 with the thirteen line segments between the transition points labelled 0 to C as hexadecimal numbers. A second ROM is used to convert the line number from the line-rate counter into codes 0 to E, as hexadecimal numbers, identifying fifteen different types of composite lines from which the complete set of waveforms can be assembled: for example, broad pulses, broad pulse followed by equalising pulse, equalising pulses, etc. The output codes identifying the line segment and the line type are used in a third ROM which defines the state of each control waveform, high or low, in each segment on each type of line. For some line types, the presence of the colour burst depends on whether the first and second or third and fourth fields are being generated. Because of this, the V-axis switch waveform, which has the opposite sense on corresponding lines of alternate pictures, is used in the third ROM to control the generation of the burst blanking sequence.

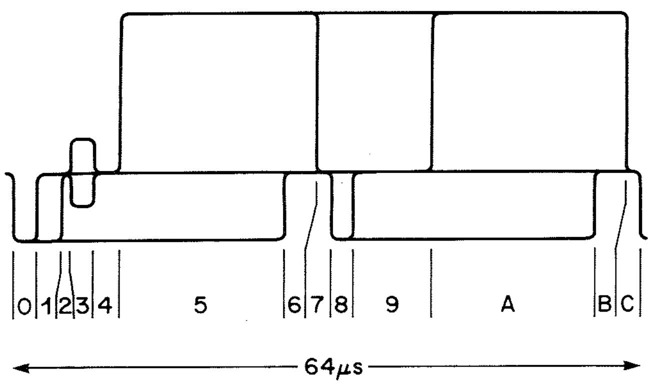


Fig. 16 – Line segment positions for mixed syncs, mixed blanking and burst gate in 625/50 PAL.

While the two-level control waveforms generated by the third ROM in Fig. 15(b) identify the positions of the edges to the nearest sample period, the sample values representing each edge have to be chosen to give an appropriate risetime and shape. A suitable shape of edge characteristic is obtained by integrating a raised-cosine impulse such as the 'T' pulse shown in Fig. 17(a). This impulse and the edge produced from it, Fig. 17(b), have the property[28] that there is virtually no signal energy beyond a frequency f where:

$$f = \frac{1}{T}$$

Therefore this characteristic provides a fast risetime, without ringing, within a well-defined bandwidth. The risetime of the edge between the 10 and 90% points is also approximately* equal to T. So, appropriate sample values can be chosen for the synchronising pulse edges (nominal risetime 0.25 μs) and for the burst envelope (0.3 μs). These values can then be read from a small ROM by a counter triggered by edges in the two-level control waveforms produced by the circuit of Fig. 15(b). Similarly, values from a 0.3 μs edge can be used as coefficients in a ROM multiplier for the insertion of blanking edges.

* More accurately, the edge duration, 10 to 90%, is equal to $0.964T$.

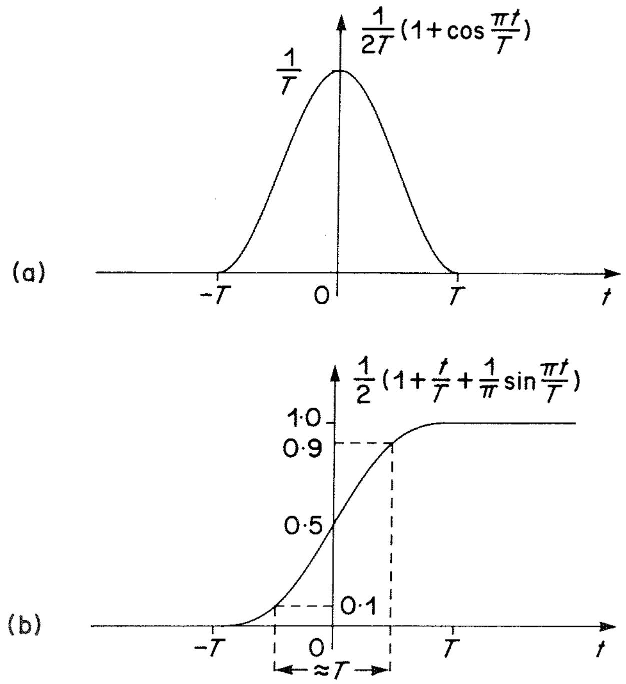


Fig. 17 – Band-limited edge generation: (a) the 'T' pulse and (b) the edge formed by integrating a 'T' pulse.

3.2.4. Filter implementation

The Gaussian U and V low-pass filters used in a PAL coder can be produced conveniently using symmetrical transversal filters with an odd number of taps. A suitably accurate response can be obtained with a nine-tap filter using the five coefficients shown in Table 1. These coefficients were derived by choosing a characteristic in the middle of the template of Fig. 8 and calculating the response value at five equi-spaced frequencies from 0 to 6.75 MHz. These values were used to solve five simultaneous equations to yield the coefficients.[29]

$a_0 = 0.3153$	$a_2 = 0.0903$	$a_4 = 0.0021$
$a_1 = 0.2308$	$a_3 = 0.0191$	

Table 1 : Coefficient values for a nine-tap transversal filter giving a Gaussian characteristic

The ladder structure of the digital transversal filters used is shown in Fig. 18. All the weighted values of a single input sample are produced simultaneously by the coefficient multipliers a_0 to a_4. These values are then added into the delay line registers so that the taps are arranged in the correct sequence. This arrangement has the advantage that the delay line elements R also act as pipeline delays for the adders. Also, if the multipliers are implemented as ROM look-up tables, the coefficient values are not subject to quantising distortion (only the products are quantised) so that the filter accurately reproduces the designed filter characteristic. Filters with a more rapid rate of cut require more coefficients and can be produced by adding more of the modular stages. Quantising noise can be maintained at a suitably low level by using ten bits for the input words (1024-word ROMS) and twelve bits throughout the adder ladder. The use of two's complement signal coding avoids the need to adjust signal offsets during the filtering process.

3.3. NTSC encoding

The NTSC system, although based on suppressed-carrier amplitude modulation of two orthogonal subcarriers, has two main areas of difference from PAL. First, there is no frequency offset between the subcarriers and secondly I (In-phase) and Q (Quadrature) colour difference signals are used, instead of the U and V signals of PAL. Thus the NTSC chrominance signal can be represented mathematically by the expression:

$$Q \sin \omega t + I \cos \omega t$$

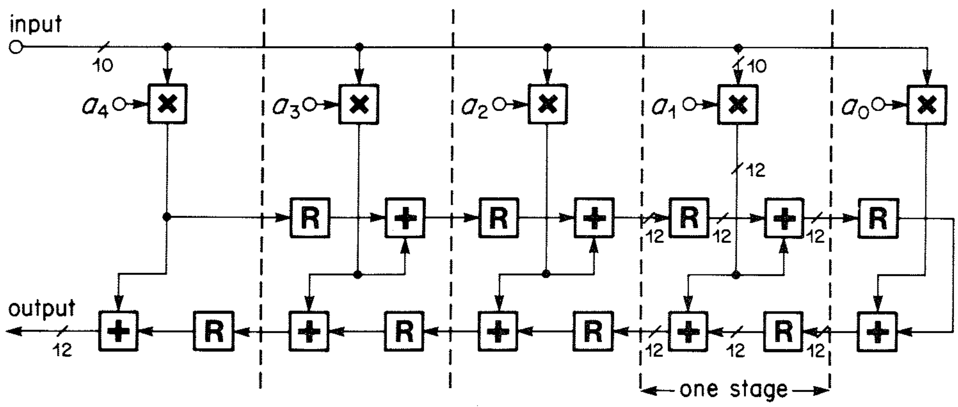


Fig. 18 – A modular transversal filter arrangement consisting of ROM multipliers, adders and registers suitable for use as a low-pass filter in a digital PAL coder.

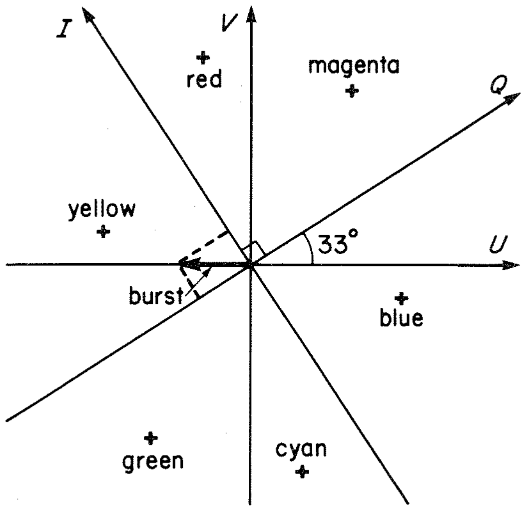


Fig. 19 – Positions of the I and Q modulation axes of the NTSC system relative to U and V, the colour primaries and their complements, and the NTSC colour burst.

where $$\omega = 2\pi f_{sc}$$

and for NTSC signals

$$f_{sc} = 3.579545\,\text{MHz}.$$

Because there is no frequency offset between the subcarriers, differential phase distortion produces static hue errors which cannot be removed by filtering. Also, sideband asymmetry results in hue errors on colour transitions. This makes the NTSC signal much less robust than PAL.

The NTSC system is based on the principle that significantly lower resolution is acceptable for some colours, notably for blue/magenta and yellow/green. So the Q and I colour difference signal axes are rotated by 33° from U and V as shown in Fig. 19 and have the following relationships to U and V:

$$Q = U\cos 33^\circ + V\sin 33^\circ$$

and $$I = V\cos 33^\circ - U\sin 33^\circ$$

This rotation aligns the Q signal with the axis of colours least affected by a loss of resolution, so minimising the impairment caused by encoding the Q signal with a low bandwidth.

In practice, with the arrangements of Fig. 10, the I and Q signals are produced either by matrixing directly from $B - Y$ or $R - Y$ signals, or from C_R and C_B signals, using the additional factors given in Section 3.2.2. These matrix arrangements are shown in Fig. 20.

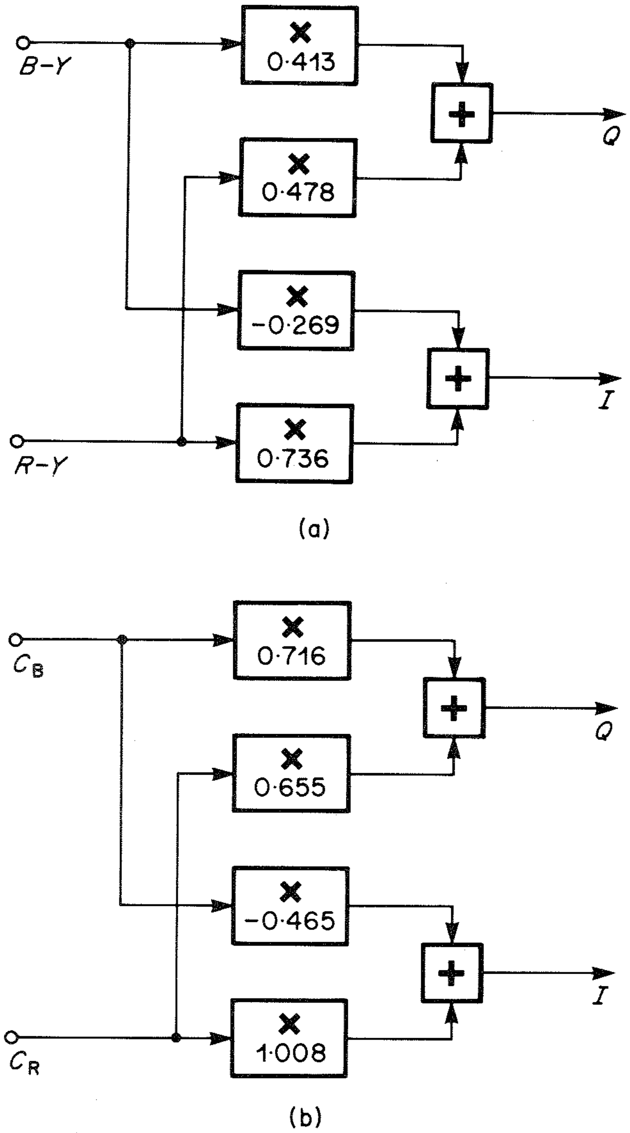


Fig. 20 – Matrix circuits for producing I and Q colour difference signals: (a) from $B - Y$ and $R - Y$ and (b) from C_R and C_B.

As the signal bandwidth of the NTSC system is only 4.2 MHz, the spectrum space available for the chrominance components is relatively limited. In particular, making the subcarrier a fine, high frequency pattern to reduce its visibility results in a significant loss of the upper chrominance sidebands. If both colour difference components were wideband, the asymmetry would cause crosstalk between the two signals. However, as the Q signal is low bandwidth, it remains a double sideband signal and causes no crosstalk. The higher frequencies from the wideband I signal, on the other hand, do cross into the Q signal, but at higher frequencies than the real Q frequencies. The I-Q crosstalk components can therefore be removed by low-pass filtering the demodulated Q signal.

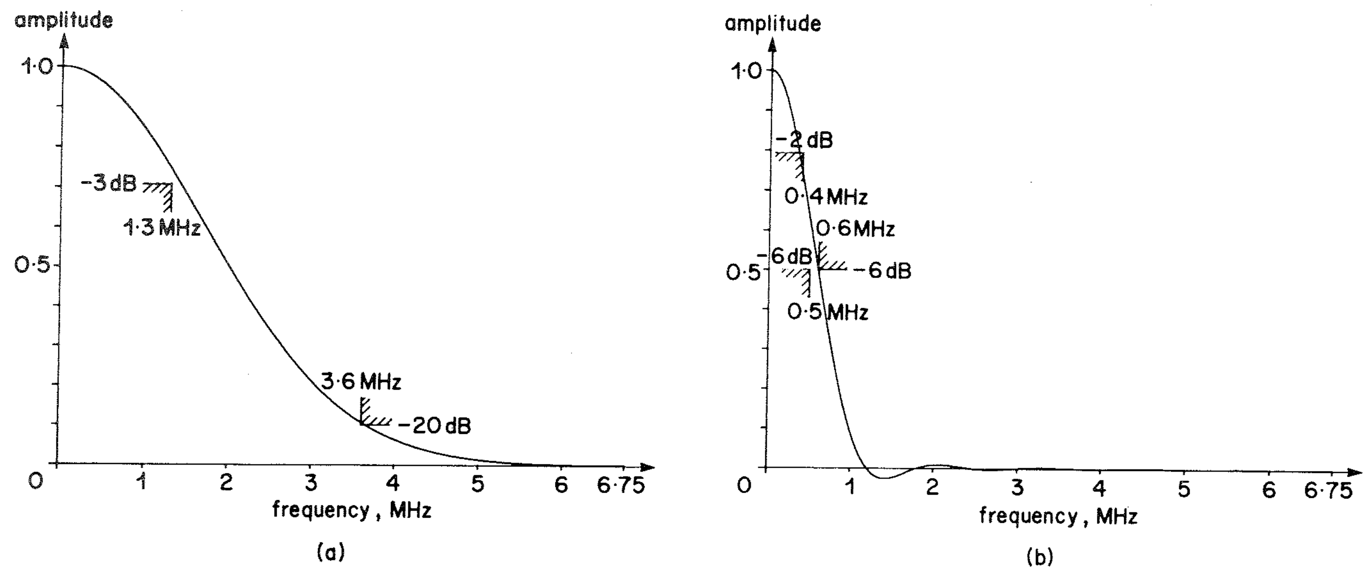


Fig. 21 – Templates for NTSC coder low-pass filters: (a) the template for the I channel showing, as an example, the characteristic produced by the coefficients of Table 1 and (b) the template for the Q channel with the characteristic produced by a 23-tap approximation using the coefficients of Table 2.

The specifications[21] for I and Q coder low-pass filters are shown in Fig. 21. While the I characteristic is similar to that of the Gaussian U and V filters used for PAL, the Q filter bandwidth is only about 0.5 MHz and has a much sharper roll-off. Although a fair approximation to the Q filter specification can be produced with a 23-tap filter using the coefficients shown in Table 2, possibly twice as many taps as this would be needed to match the template accurately. Additional delays are needed in the I and Y signal channels to compensate for the extra delay of the Q filter.

$a_0 = 0.0889$	$a_4 = 0.0639$	$a_8 = 0.0171$
$a_1 = 0.0872$	$a_5 = 0.0522$	$a_9 = 0.0085$
$a_2 = 0.0821$	$a_6 = 0.0398$	$a_{10} = 0.0027$
$a_3 = 0.0742$	$a_7 = 0.0278$	$a_{11} = 0.0002$

Table 2 : Coefficient values for a 23-tap approximation to the NTSC Q-filter specification

As the NTSC colour burst lies on the $-U$ axis, as shown in Fig. 19, this could be generated as proportions of I and Q. However, in a digital coder, it is perhaps more convenient to insert the envelope purely in the Q signal and to alter the subcarrier reference phase by the addition of a 33° offset during the burst. Also phase switching of the V subcarrier is not required for NTSC.

The final feature of the NTSC signal is included by low-pass filtering the encoded signal to 4.2 MHz to remove part of the upper sideband of the I signal. Where access to a digital NTSC signal is not required, this low-pass action can be produced by using a 4.2 MHz filter in the digital-to-analogue converter.

A considerable simplification of a digital NTSC coder can be made by ignoring the need for a narrow bandwidth Q filter. Instead, wideband Gaussian filters can be used in both channels, as in PAL. If both filters are the same, the matrix used to rotate the modulation axes to I and Q is not needed, so that the U and V signals can be modulated directly. Also, the colour burst can be generated directly in the U channel without the need to offset the subcarrier by 33° and there is no need for a 4.2 MHz output low-pass filter.

In the studio this so-called 'broadband' NTSC remains compatible with normal NTSC and provides improved quality on colour transitions. When bandlimited to 4.2 MHz as may occur in a transmitter, the upper sideband is lost. This results in crosstalk between the Q and I channels in the decoder on the high frequency colour components. This shows up mostly as static hue errors on edges, although the effect is unlikely to be very serious. This is because the hue is incorrect on edges in normal NTSC due to the loss of high frequency Q signals. The broadband approach could, therefore, be included as a simple modification to a digital PAL coder in circumstances where the full extra complexity of NTSC coding is not required.

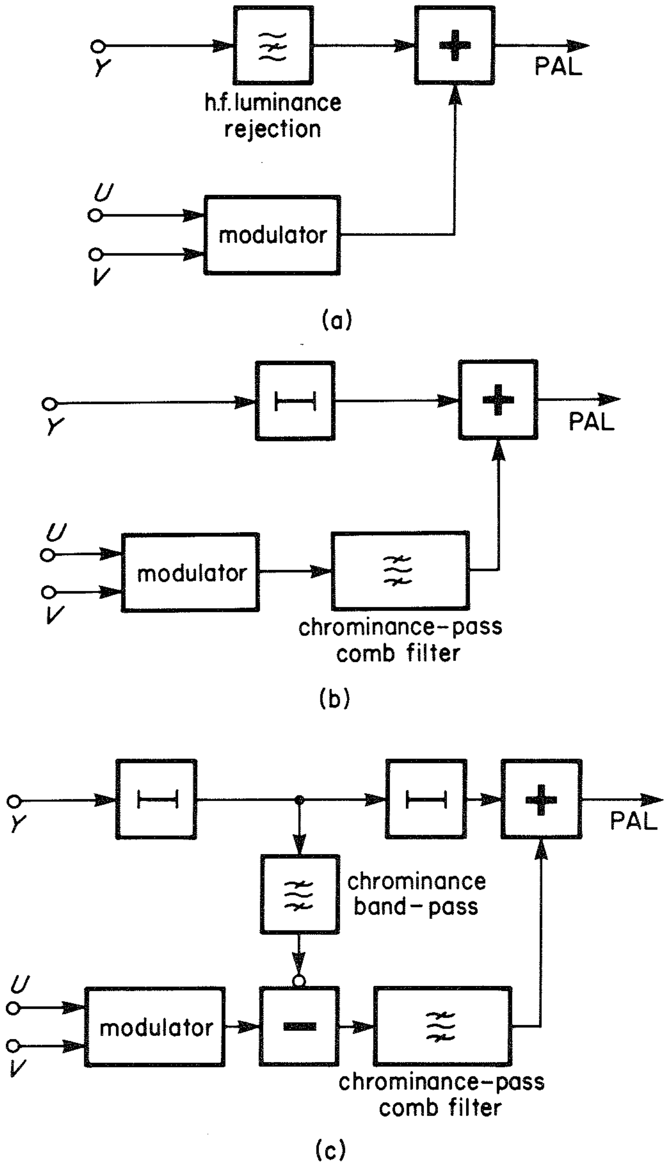


Fig. 22 – Clean PAL coding systems: (a) the removal of h.f. luminance to reduce cross-colour, (b) the removal of vertical/temporal chrominance detail to reduce cross-luminance, and (c) a complementary clean PAL coder arrangement.

3.4. Clean coding systems

The only filters present in a conventional PAL coder are the Gaussian low-pass filters used to shape the spectrum of the colour difference signals. Because of this, there is considerable spectral overlap between the luminance and chrominance components, which, in general, makes it impossible to separate the signals completely by filtering at the decoder. However, in both PAL and NTSC, additional filtering can be included at the coder to reduce the severity of subsequent cross-colour and cross-luminance impairments.

Cross-colour in the decoder can be reduced by removing some of the high frequency luminance in the coder. For PAL, the simplest example of this is the inclusion of a luminance notch filter, as shown in Fig. 22(a), which, although reasonably effective when conventional decoders are used,[30] reduces the potential for improvements in the decoder.

With the stability of digital circuitry, more complicated vertical and temporal comb filtering techniques[16,31] can be employed in the coder. Such techniques allow fine, moving luminance, which produces the more noticeable coarse, stationary cross-colour, to be removed while retaining the full static luminance resolution. Similar techniques can be applied in the chrominance channel, as in Fig. 22(b), to reduce cross-luminance. This can be achieved by restricting the vertical and temporal chrominance sidebands, although if too severely applied, the chrominance information can become blurred.

Many rearrangements of these clean coder circuits are possible, in some of which the colour difference signals are filtered before modulation. A particularly efficient rearrangement is shown in Fig. 22(c), in which a single chrominance-pass comb filter is used to provide cross-colour and cross-luminance suppression. With 'ideal' rectangular filters, such an arrangement would ensure that luminance and chrominance were kept perfectly separate in the spectrum. However, the use of rectangular filters is ruled out by the severe ringing which would occur at abrupt transitions, as well as by the high cost of sharp-cut filters.

A more satisfactory method of ensuring perfect separation of luminance and chrominance is to use the Weston clean coding and decoding technique.[10,12,14] This produces a compatible variant of PAL in which the two chrominance signals are carried on one phase of subcarrier, whilst high frequency luminance is carried on the orthogonal phase.* Because of this, the luminance and chrominance signals can be separated perfectly by synchronous demodulation.

Although both additional coder filtering and matched coder-decoder combinations can improve the picture quality obtained from PAL signals, the introduction of such systems into a conventional PAL environment does present problems. With matched systems that require both modified coders and decoders, such as Weston clean PAL, the improvements for viewers with conventional decoders are small and are sometimes offset by the introduction

* The Weston system was first explained in these terms by N.D. Wells.

of additional impairments. Also such systems require expensive receivers and compatibility with archival, conventional PAL tapes has to be considered. On the other hand, in systems in which modifications are confined to the coder, the improvements are less spectacular, but still beneficial from a picture quality viewpoint. However, the large number of PAL coders used and the studio timing difficulties caused by the introduction of extra delays in the comb filters currently remain as considerable drawbacks. In the future, though, when large areas have been converted to digital component signal operation, the use of complicated clean coders for the network outputs could be worthwhile, provided that compatibility with sub-Nyquist sampled distribution links can be maintained.

4. Colour decoding

4.1. Decoder configurations

While the luminance and chrominance components in a conventional colour encoder are combined by simply adding the two signals together, separating the signals again in a decoder is much more difficult. More complicated separation techniques can be used in digital decoders, however, because of their stable performance and consequent freedom from unexpected signal distortions.

The main processes of a colour decoder consist of separating the luminance and chrominance, and demodulating the chrominance to produce two

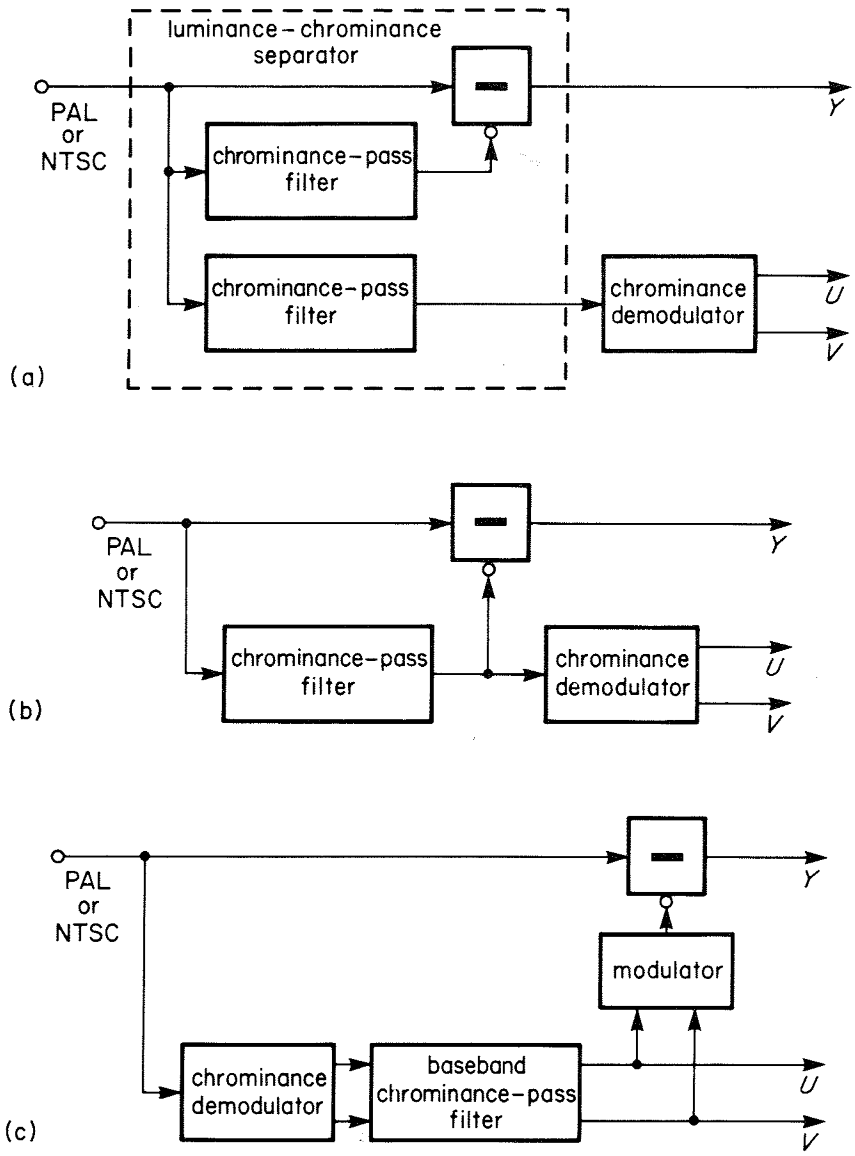


Fig. 23 – Decoder configurations: (a) with non-complementary luminance and chrominance separation filters, (b) with complementary filtering and (c) with complementary filtering of the baseband colour difference signals.

colour difference signals. When comb filters are used, it is often convenient to obtain the luminance components by first separating a chrominance signal and using this to cancel appropriately delayed chrominance components from the input signal, as shown in Fig. 23(a). Although the use of different chrominance pass characteristics is often advantageous, it is sometimes possible to simplify the circuit to use only one filter, as in Fig. 23(b). A further rearrangement which can produce equivalent results is shown in Fig. 23(c). In this case the chrominance is demodulated first and then the unwanted luminance components (cross-colour) are suppressed with baseband colour difference filters. Chrominance components in the input signal (cross-luminance) are cancelled using a remodulated version of the filtered colour difference signals.

Many different comb filters have been used to separate chrominance and luminance signals,[16,31] and work is still continuing to find the optimum method. In contrast, the demodulation techniques described in this Section can be applied satisfactorily in any of the arrangements shown in Fig. 23. Also described here are techniques such as PAL modification which is used in chrominance-luminance separation filters, but has much in common with synchronous demodulation.

4.2. Line-locked sampling of composite signals

4.2.1. Effects of quantising distortion

When a television signal is encoded as a stream of binary numbers, the process of quantisation distorts the signal to produce low-level harmonics of the signal frequencies. The quantisation errors are partly the result of the fundamental accuracy of coding (usually at least eight bits are used) and subsequent processing and partly the result of instrumental errors in a–d and d–a conversion equipment. When coupled with the effect of sampling, the harmonic components can be converted to low frequencies which may be visible as beat patterns on the picture. These are more noticeable on composite colour signals when the high frequency chrominance components are demodulated to form baseband colour difference signals.

The form of the beat patterns depends on both the signals and the sampling frequency. Special cases occur with subcarrier-locked sampling and signals representing areas of constant colour, for example, colour bars. Then the distortion results in a constant hue error which is much less noticeable. With the same signal and line-locked sampling, the beat pattern visibility is not dissimilar to that at other frequencies (non-line-locked). It was partly for this reason that subcarrier-locked sampling was regarded as superior to line-locked sampling.[32] However, this result reflects the exactly constant levels of the colour bar signal and is not obtained when other signals are considered. In fact, with gradually changing signals, such as a constant hue on a luminance sawtooth signal, the beat patterns produced by subcarrier-locked sampling are somewhat more noticeable than those produced by line-locked sampling. So far as the accuracy of quantisation is concerned, therefore, there is little to choose between the two.

Fortunately with modern a–d conversion equipment and eight-bit quantisation, the level of beat patterns introduced by digital encoding is generally below the level of perception.[3,33] Also the use of low-level dither signals[3] in the a–d conversion process can disguise the beat patterns for any sampling frequency by converting them to random noise. Similar results are obtained when digitising NTSC signals.

4.2.2. Sampling frequency stabilisation

If the comb filters of a digital decoder contain field or picture delays, then the pulses used to sample the composite colour signal have to be particularly stable. With such a long delay between storing and reading the samples, even a small clock frequency error can produce a variation of delay that is a significant proportion of a subcarrier period. Such delay variations would destroy the effectiveness of a comb filter for separating chrominance and luminance signals.

It has proved much easier, in practice, to obtain a stable sampling frequency with subcarrier-locked sampling, in which the colour burst is used as the reference, than with line-locked sampling, which uses the line pulse falling edge as the reference. This might be ascribed to the greater timing information content of the burst, although this has been calculated* as being between only twice and seven-times that of the sync. edge, depending on whether the noise spectrum of the video signal is assumed to be triangular or flat. A more significant factor is the use of analogue slicing techniques, which are particularly susceptible to noise amplification, in obtaining the line pulse reference. These are less effective than the synchronous demodulators used in phase comparators for subcarrier-locked sampling.

A more effective phase comparator for line-locked sampling can be produced using digital circuitry immediately following the a–d converter,

* This calculation was made by T.A. Moore.

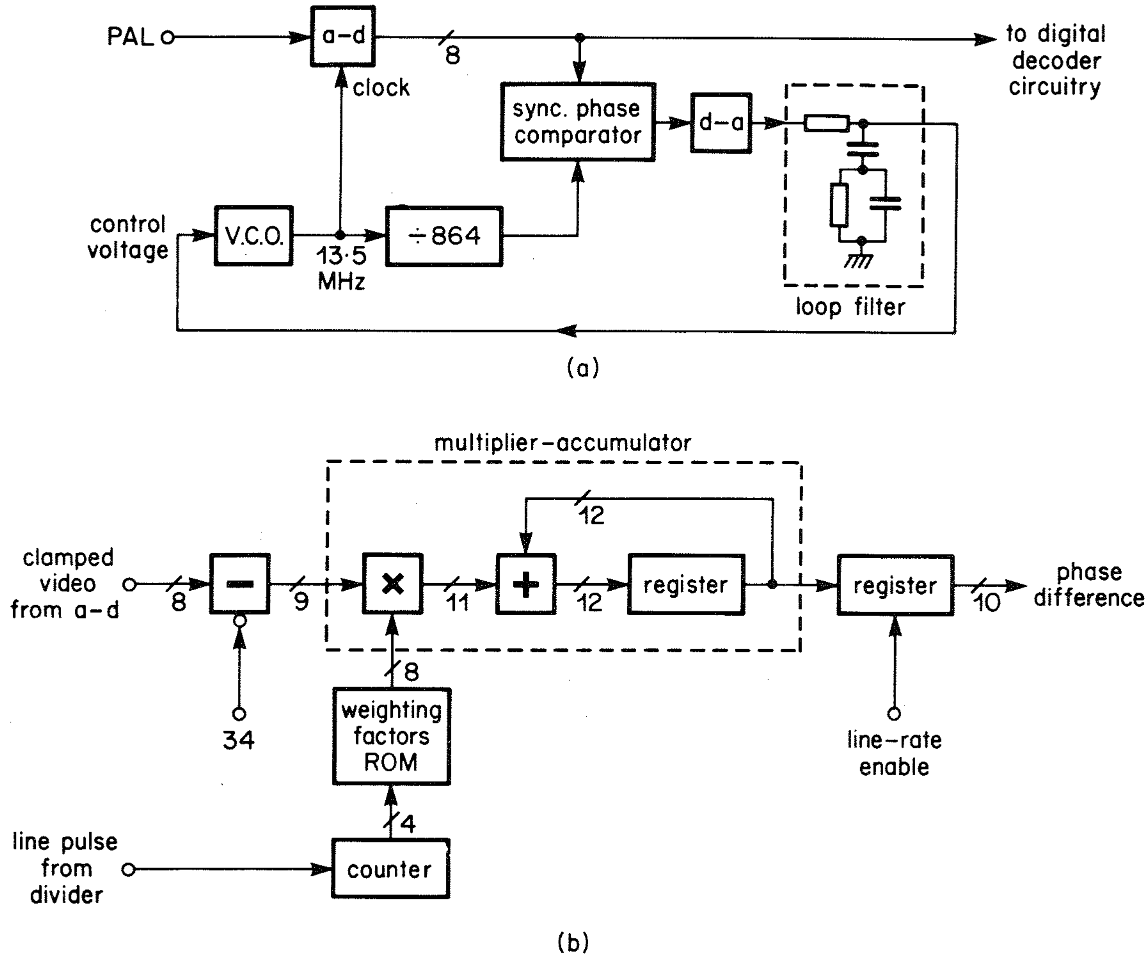


Fig. 24 – (a) A digital loop for stabilisation of a line-locked sampling oscillator. (b) The sync. phase comparator shown in more detail.

as shown in Fig. 24. The digital sample values describing the sync. pulse falling edge are compared in Fig. 24(a) with a line pulse produced by division from the clock frequency. The resulting control signal, changing at line rate, is converted to analogue form to adjust the frequency of a 13.5 MHz crystal oscillator using a conventional loop filter.[34]

The phase comparator, shown in more detail in Fig. 24(b), requires the black level of the video signal to be clamped in the a–d converter, usually at level 64 for PAL. Then with a standard level signal and the converter gain set to 5 mV/step, the coding level corresponding to the line timing reference point is 34 as shown in Fig. 25(a). Because of this, 34 is subtracted from each video sample so that the waveform value at the timing reference point is zero. A control waveform could be produced by sampling this waveform with the line pulse from the 864 divider. This would give zero error when the divider line pulse was coincident with the timing reference point. However, a more effective control signal can be obtained by summing a series of weighted samples from the region of the sync. edge using the multiplier-accumulator shown in Fig. 24(b). This arrangement uses more of the timing information from the sync. edge and can be used to suppress noise. Accumulating the weighted samples is equivalent to a transversal filter with an impulse response given by the series of weighting factors. The optimum suppression of noise can be obtained by relating the frequency characteristic of this filter to the spectral content of the edge. Also the filter should have zero response at subcarrier frequency to ensure that chrominance components extending below black level are suppressed. The frequency characteristic of such a filter is shown in Fig. 26, which is produced by the coefficients of Table 3.

$a_0 = 0.265$	$a_3 = 0.044$	$a_6 = -0.011$
$a_1 = 0.224$	$a_4 = -0.004$	$a_7 = -0.003$
$a_2 = 0.132$	$a_5 = -0.016$	

Table 3 : Weighting factors used in the line phase comparator

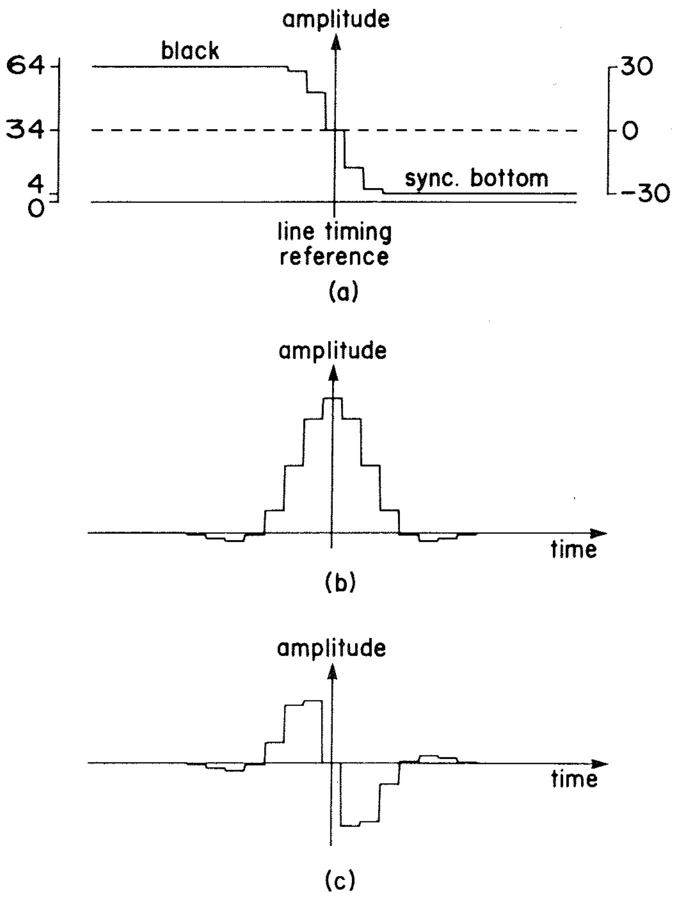


Fig. 25 – Waveforms in the digital line phase comparator: (a) the line sync. edge, (b) the series of weighting factors and (c) the weighted sync. edge samples.

To perform the filtering, the weighting factors are read from a read-only memory by a counter triggered from the 864 divider line pulse. This produces the waveform of Fig. 25(b) at the multiplier input which, when multiplied by the sync. edge, produces the waveform of Fig. 25(c). When the central weighting factor is coincident with the line timing reference, this waveform integrates to zero. Distortion of the sync. edge causing a loss of anti-symmetry can alter the locking position, but the error is usually negligible.

Although the loop will lock with this phase detector alone, the lock-up time can be rather slower than necessary if the divider phase has a large offset from the incoming sync. phase. This can be avoided by initially using a conventional sync. separator to force the 864 divider into approximately the correct phase before transferring control to the phase detector.

The phase detector of Fig. 24(b) uses the assumption that the input signal will be at standard level. If the signal level is not standard, the locking position will vary with the sync. height until at half amplitude the circuit will fail to lock. This can be avoided by a small modification which involves subtracting 64 from the incoming signal instead of 34. The filter therefore accumulates a negative value at the half amplitude position of the sync. edge which is then offset by subsequently accumulating additional values taken from the sync. bottom. Locking at half the sync. height is then assured by weighting the samples taken at the sync. bottom by coefficients with half the amplitude and the opposite sign to those used on the sync. edge.

Implementing the line phase detection circuits after the a–d converter eliminates the possibility of noise pick-up associated with analogue sync. slicing techniques. This provides sufficient sampling clock stability to avoid any significant variations in field and picture delay comb filters. Also, because the a–d converter is included in the control loop, a specified sampling pulse to sync. phase is ensured in the digital signal. Thus the registration of the picture information on the sampling grid is reproducible.

Although controlled by the phase-locked loop, the stability of the sampling pulses is in part dependent on the basic stability of the sampling clock oscillator. For this reason, a crystal oscillator is normally used. However, the locking range of a simple crystal oscillator is insufficient to accommodate the full range of non-standard line frequencies. While sufficient range for all non-standard signals could be obtained with an *L-C* oscillator, adequate basic stability can be difficult to achieve.

4.3. PAL chrominance demodulation

As described in Section 3, the PAL chrominance signal can be represented by the expression:

$$U \sin \omega t \pm V \cos \omega t$$

with the sign of the V component alternating from one line to the next. The colour information can be demodulated by multiplying the signal by appropriately phased subcarrier frequency sine waves and low-pass filtering the resulting products, as shown in Fig. 27(a). This is known as synchronous demodulation. U is obtained by multiplying by $2 \sin \omega t$ and V is obtained by multiplying by $\pm 2 \cos \omega t$, so that:

$$(U \sin \omega t \pm V \cos \omega t) 2 \sin \omega t$$
$$= U - U \cos 2\omega t \pm V \sin 2\omega t$$

and

$$(U \sin \omega t \pm V \cos \omega t)(\pm 2 \cos \omega t)$$
$$= V \pm U \sin 2\omega t + V \cos 2\omega t$$

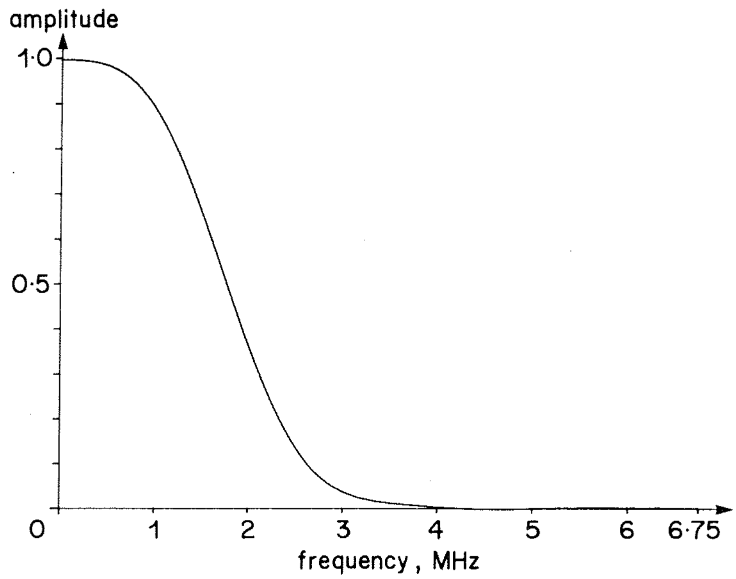


Fig. 26 – Low-pass characteristic corresponding to the filtering action of the line-phase comparator.

In each case, the twice subcarrier frequency (2ω) terms are removed by the low-pass filters, leaving only the U and V signals at the demodulator outputs. Using a switched subcarrier waveform in the V channel removes the PAL switch modulation as part of the same process.

If the local reference subcarrier phase were incorrect, as could result from differential phase distortion, then the line-to-line pattern known as Hanover bars would result. This can be shown by including a reference phase error θ making the demodulating subcarriers $2\sin(\omega t - \theta)$ and $\pm 2\cos(\omega t - \theta)$, so that now:

$$(U \sin \omega t \pm V \cos \omega t) 2 \sin(\omega t - \theta)$$

and

$$(U \sin \omega t \pm V \cos \omega t)(\pm 2 \cos(\omega t - \theta))$$

produce

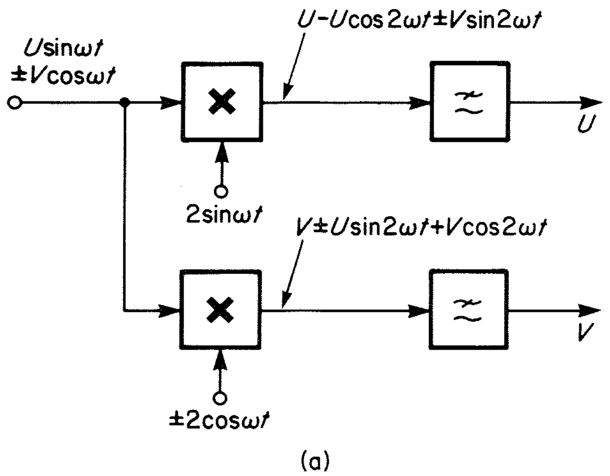
$$U \cos \theta \mp V \sin \theta$$

and

$$V \cos \theta \pm U \sin \theta$$

after low-pass filtering. In areas of constant colour, averaging equal contributions from odd and even lines by, for example, averaging across a line delay, in each case cancels the alternating crosstalk component, leaving only a desaturation of the true component by the factor $\cos \theta$.

In a conventional analogue demodulator, the incoming reference burst is used to lock a subcarrier frequency crystal oscillator which includes its own phase detector, as shown in Fig. 27(b). Alternate bursts pull the oscillator towards 135° and 225° so that, by including a long time constant in the loop filter, the reference oscillator takes up the mean phase of 180°. The error signal of the phase detector then provides the alternating V switch square wave directly.



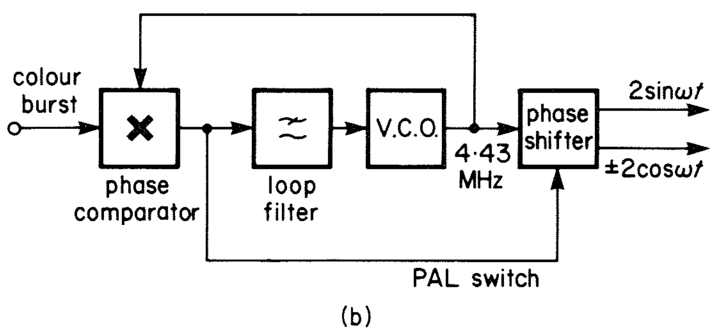


Fig. 27 – PAL chrominance demodulation: (a) signals in a PAL demodulator and (b) the separate phase detector arrangement used in a conventional analogue demodulator.

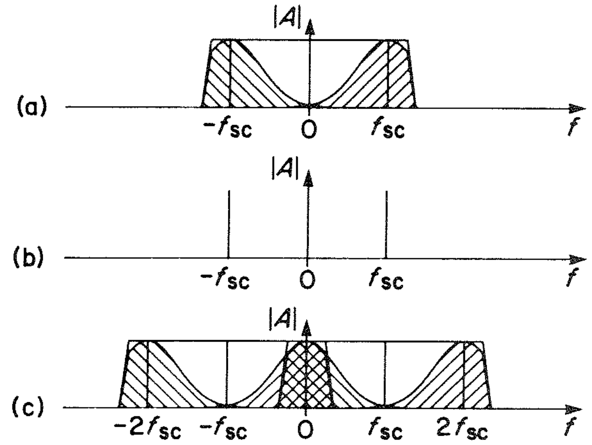


Fig. 28 – Frequency spectra for chrominance demodulation: (a) the spectrum of an analogue video signal with chrominance information in the shaded region around the colour subcarrier frequency, (b) the spectrum of a subcarrier frequency sine wave, and (c) the result of convolving (a) and (b).

The demodulation process is shown in spectral terms in Fig. 28. Fig. 28(a) represents the baseband spectrum of an analogue video signal with chrominance information in the shaded region around the colour subcarrier frequency, whilst Fig. 28(b) represents the spectrum of the subcarrier frequency sinewave used for demodulation. Convolution of these two spectra (equivalent to multiplication in the time domain) produces the spectrum of Fig. 28(c) in which the baseband video spectrum has been shifted to be centred on f_{sc} and $-f_{sc}$. Therefore, the chrominance information which now appears as a baseband signal can be separated from the low frequency luminance centred on f_{sc} and the chrominance centred on $2f_{sc}$ by a low-pass filter.

The frequency characteristic of the demodulation low-pass filters is a compromise between several factors. To extract all of the transmitted chrominance resolution requires a wide passband, which would increase the amount of cross-colour because a greater range of luminance frequencies would be included. Also, above 1.07 MHz (see Section 3.1), any loss of the upper chrominance sidebands would introduce ringing and *U-V* crosstalk. Filters with a sharp-cut at 1.07 MHz would maximise the retention of chrominance while avoiding *U-V* crosstalk on edges and much of the cross-colour. However, such filters would accentuate ringing on chrominance edges and for this reason slow roll-off filters are normally used. For professional decoders, these have a relatively wide passband, similar to the colour difference filters used in PAL coders (Fig. 8). Decoders used in domestic receivers, however, tend to have a narrower bandwidth, thus producing poorer resolution, but reducing cross-colour, ringing and *U-V* crosstalk on edges.

4.3.1. Digital demodulation

In a sampled system, the demodulation process is similar to that for continuous signals except that the baseband spectra of Fig. 28 are repeated around harmonics of the sampling frequency. This is shown in Fig. 29 for a sampling frequency of 13.5 MHz which is slightly greater than three-times the PAL subcarrier frequency. While the sampled video and subcarrier spectra, Figs. 29(a) and (b), are identical to those of Fig. 28 up to half the sampling frequency, the repeated spectra of the demodulated signal, shown in Fig. 29(c), overlap to produce alias components which were not present in the corresponding analogue signal spectrum. However, with sampling frequencies above the Nyquist rate for the video signal, these alias components do not overlap with the main components of the baseband chrominance spectrum and therefore can still be removed by low-pass filtering.

In a digital decoder, the synchronous demodulators of Fig. 27(a) consist of four-quadrant digital multipliers followed by digital transversal filters. A similar filter design method can be used to that described for the coder in Section 3.2.4. However, the decoder requires a slightly sharper roll-off to ensure suppression of the sampling alias components which, with a 13.5 MHz sampling frequency, start to become significant above the 3 MHz region.

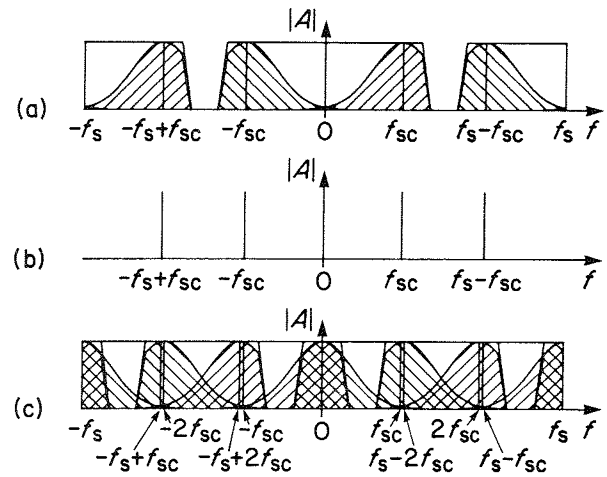


Fig. 29 – Frequency spectra for digital PAL chrominance demodulation at $f_s = 13.5$ MHz: (a) the spectrum of sampled composite video, (b) the spectrum of a sampled subcarrier sine wave, and (c) the result of convolving (a) and (b).

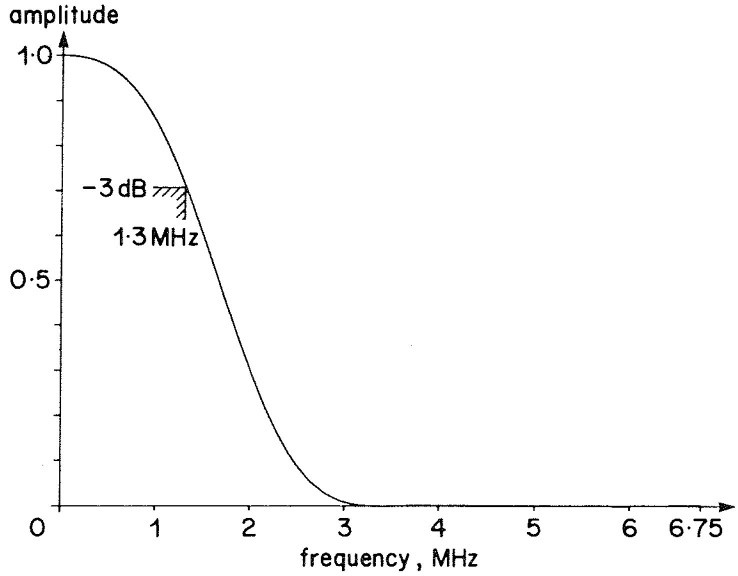


Fig. 30 – The frequency characteristic of a low-pass filter for use in a digital PAL demodulator. This is obtained from the coefficients of Table 4.

A suitable characteristic is shown in Fig. 30, which can be obtained with a 15-tap filter using the coefficients of Table 4.

$a_0 = 0.2460$	$a_3 = 0.0540$	$a_6 = -0.0103$
$a_1 = 0.2136$	$a_4 = 0.0019$	$a_7 = -0.0031$
$a_2 = 0.1356$	$a_5 = -0.0146$	

Table 4 : Coefficient values for the demodulator low-pass filter characteristic of Fig. 30

4.3.2. Subcarrier-to-burst synchronisation

The subcarrier generation process in a digital demodulator is based on the same ratio counting method already described in Section 2. However, the local phase signal generated by the ratio counters must be adjusted initially to match the reference phase of the incoming signal. Moreover, if the subcarrier and line frequencies are not in the normal relationship, then continuing adjustments of the local phase are necessary.

These adjustments are made automatically by a digital phase-locked loop, the main units of which are shown in Fig. 31. The digital loop derives a phase error signal by examining the demodulated colour burst signals produced at the *V* output of the demodulator. This avoids the separate multiplier used for burst phase detection in the conventional analogue demodulator shown in Fig. 27(b). Because of this, any comb filtering preceding the digital demodulation must be bypassed for the duration of the burst.

Before the loop is locked, the arbitrarily phased subcarriers demodulate the incoming signals,

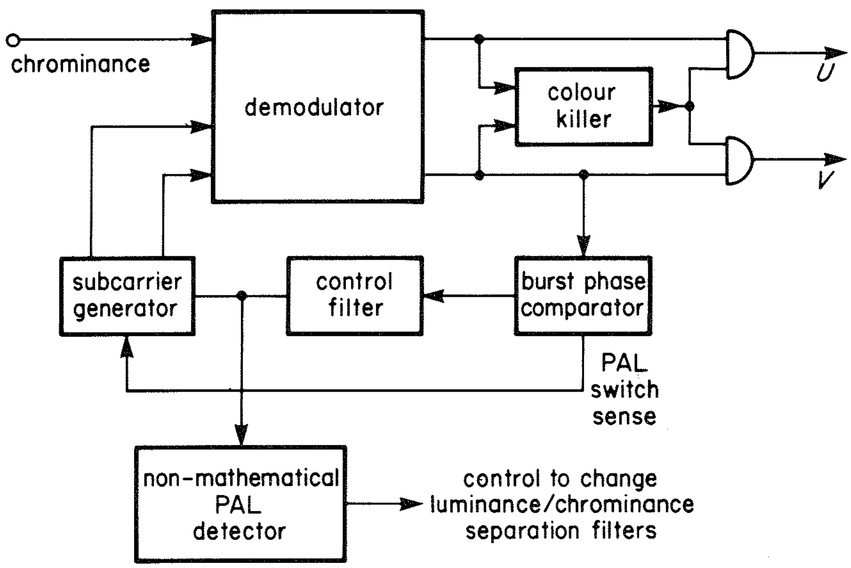


Fig. 31 – The subcarrier phase synchronisation loop in a digital decoder.

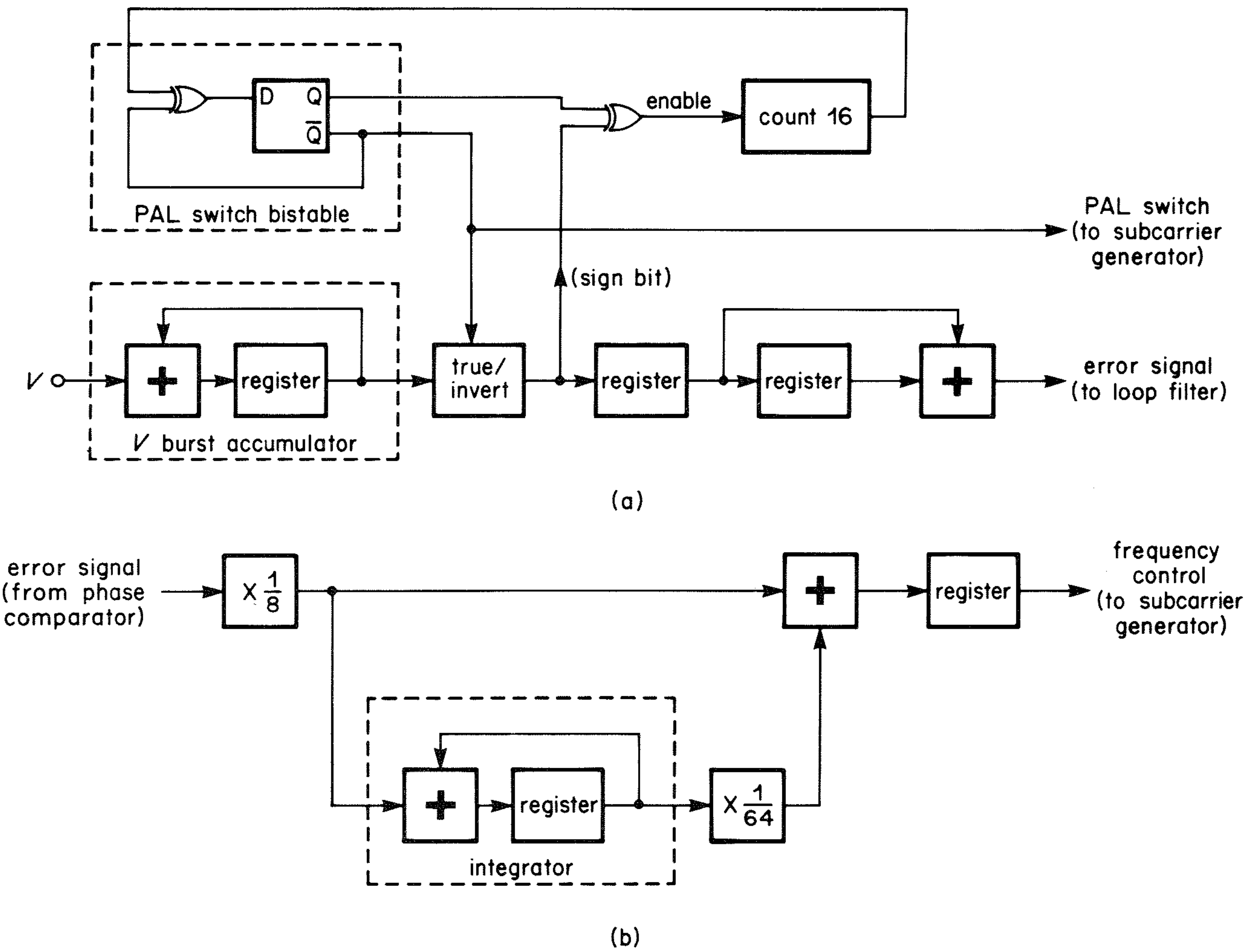


Fig. 32 – (a) The phase comparator and (b) the control filter of the digital subcarrier phase-locked loop.

including the burst. This produces signals related to the difference between incoming bursts and the local subcarrier phase. The phase comparator, shown in Fig. 32(a), is triggered by a pulse derived from the line timing reference to accumulate sixteen consecutive sample values from the nominally flat, middle region of the demodulated *V*-channel burst waveform. Although more samples could be used, this is not necessarily beneficial. This is because the edges of the burst sometimes suffer transient distortions which would tend to reduce the accuracy of phase measurement.

The phase detector also contains the local PAL switch bistable which provides the line alternating phase inversion used in generating the switched *V* subcarrier. The same signal is used to reintroduce the PAL switch into the control loop by inverting alternate values produced by the *V*-burst accumulator. The local PAL switch bistable is synchronised to the incoming signal by comparing the local switch sense with the sign of the accumulated switched burst values. If the sense is consistently incorrect for several lines, then the bistable is reset as shown in Fig. 32(a). The accumulated values from two adjacent lines are then averaged to produce the error signal. When the local subcarrier is correctly phased, the accumulated values from alternate lines cancel, so that the phase error signal is zero.

The line rate samples of the error signal produced by the phase detector are filtered to give the loop suitable stability and locking characteristics. The control filter shown in Fig. 32(b) produces a subcarrier frequency control signal consisting of proportional and integrated contributions. The gain factors shown were chosen using a computer model of the loop to give a rapid, slightly underdamped lock-up with a time constant of about ten lines. A rather slower response might be preferable for very noisy signals. Also, at the beginning and end of the burst blanking period (the nine lines without bursts in each field interval) there is a failure of cancellation which causes transients in the error signal. The rapid

response of the loop ensures that this is not a problem without the need to blank out these spurious error signals.

For non-mathematical operation, the control filter has to maintain a non-zero output to provide the appropriate frequency offset in the locally generated subcarrier. It should be noted that, when the line frequency is incorrect, the subcarrier frequency generated by the ratio counters will also require correction, even if the incoming subcarrier frequency is correct. This is because the sampling frequency, being line-locked, is also incorrect. With proportional control alone this would cause the loop to lock with a steady phase error. The inclusion of the integrator allows the filter to maintain an output without an input, so allowing the loop to be locked to non-mathematical signals in both frequency and phase. As designed, the loop can accommodate subcarrier frequency offsets of approximately ± 100 Hz. Although adequate for most purposes, this limitation is caused by number overflow and could be improved by extending the arithmetic to more bits.

The filtered error signal is added to the increments used in the lower section of the ratio counters to alter the local subcarrier frequency by a small amount. With the 16-bit counters shown in Fig. 33(a), the least significant bit corresponds to approximately 0.1 Hz change in subcarrier frequency. Whereas the correct ratios must be used in a coder, the subcarrier phase generator of a decoder could be simplified by substituting an approximate fraction with a binary denominator. Although the same number of bits would be required in the ratio counter, the modulo-q overflow would be automatic, thus eliminating the data selector and one of the frequency control adders, so producing the arrangement of Fig. 33(b). The error in frequency would be corrected automatically by the phase-locked loop adding a small offset at the frequency control input.

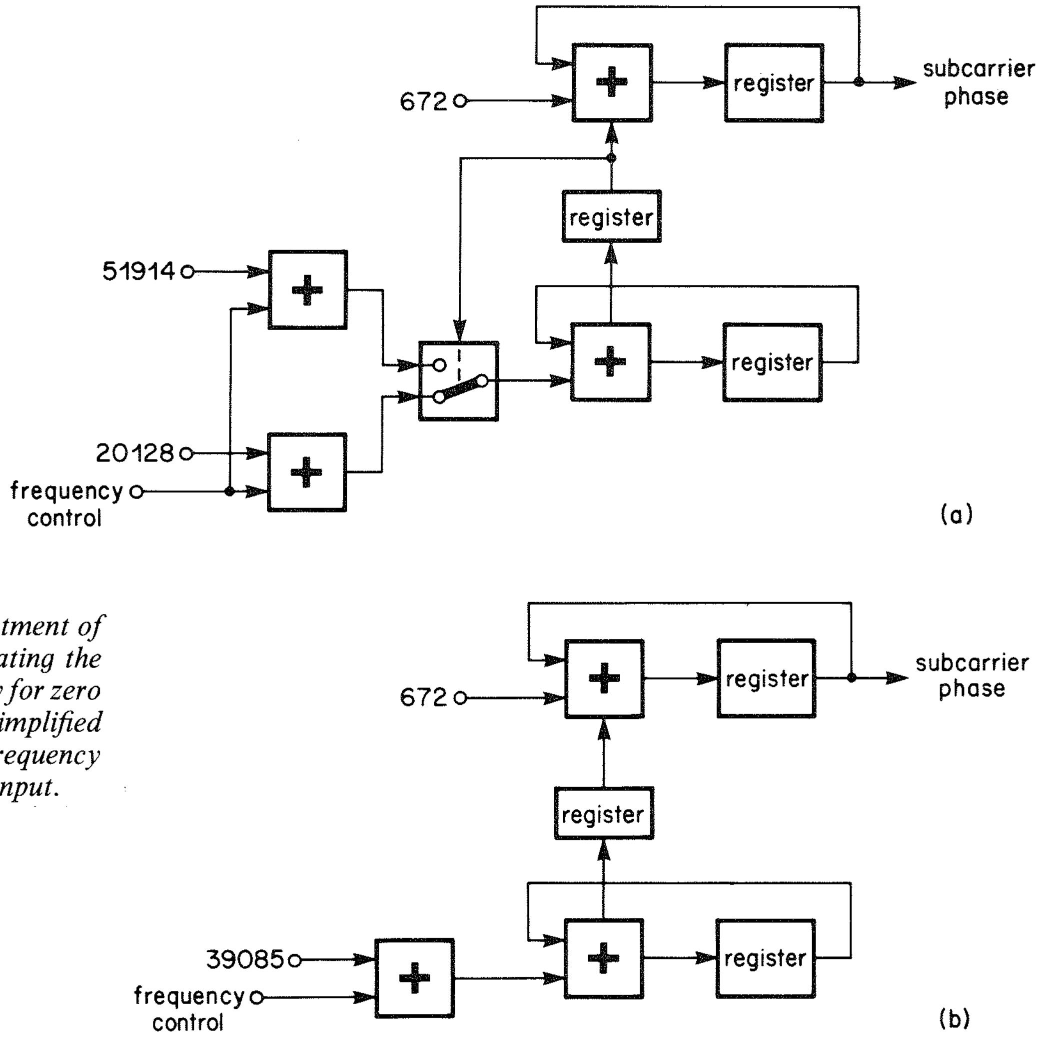
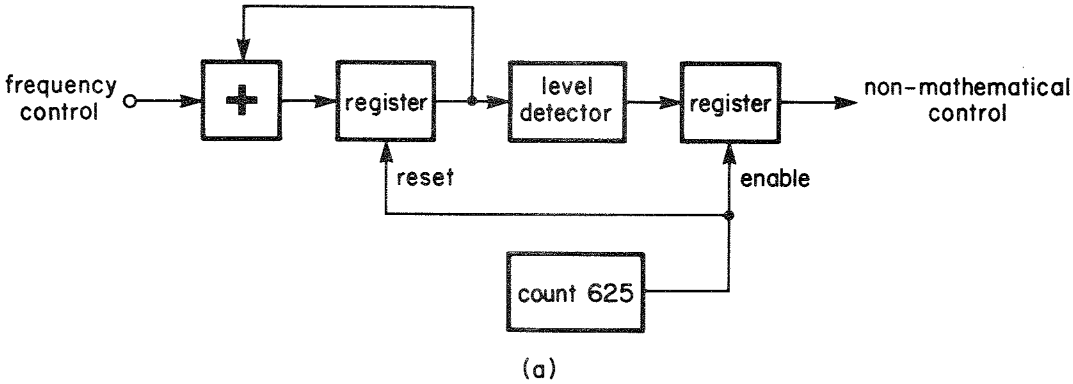


Fig. 33 – Frequency adjustment of ratio counters: (a) generating the exact subcarrier frequency for zero control input and (b) a simplified circuit which has a small frequency error for zero control input.



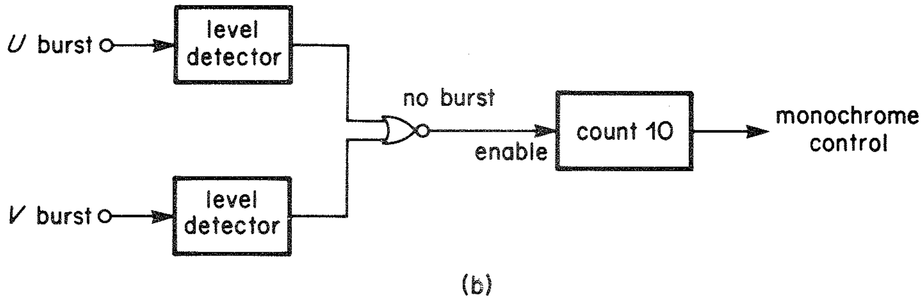


Fig. 34 – Circuit arrangements for detecting (a) non-mathematical PAL and (b) monochrome signals in a digital PAL demodulator.

A non-mathematical PAL input signal can be detected as shown in Fig. 34(a) by integrating the signal at the frequency control input of the subcarrier phase generator for a picture period. When the correct ratios are used, this signal integrates to zero for mathematical signals, or to a known value if an approximate ratio is used. Thus, a level comparator can be used to detect if the value is significantly different from that for a mathematical signal. Integration for a picture period is used because picture content and noise can produce small line-to-line variations in the control signal. With a non-mathematical error of more than about 1 Hz in the subcarrier frequency, some field and picture-based comb filters do not completely suppress the subcarrier components. The result is cross-luminance and hue errors in coloured areas of the picture. Under these circumstances, it is possible to use the non-mathematical detection signal to switch to a line-based comb filter method, which is substantially unaffected by the frequency error.

The demodulation system of Fig. 31 also includes a means of detecting the absence of the colour burst in monochrome signals. This circuit, shown in detail in Fig. 34(b), samples the demodulated *U* and *V* signals in the middle of each colour burst period. If neither channel includes a burst for more than the nine lines of burst blanking, the signal is recognised as being monochrome. The control signal produced can then be used to suppress the 'colour difference' signals, thus avoiding cross-colour. Also since there will be no subcarrier to suppress, all luminance filtering can be removed so that the full signal bandwidth is displayed.

These features of the line-locked sampling demodulator method allow the full range of colour signals to be accommodated without undue complication. Also the method avoids passing on, through the sampling frequency, any signal frequency defects which might cause complications in a subsequent digital components studio system. Any difference of line frequency or phase can then be adjusted to conform to the local reference using a digital component signal synchroniser.

4.3.3. Signal relationships in decoders

The conversion and scaling operations required in a digital PAL decoder are, in general, the reverse of those described for PAL encoding in Section 3.2.2. Fig. 35 shows the options of converting either to analogue *RGB* and syncs. or digital 4:2:2 multiplexed component signals.

Unweighted demodulation of chrominance produces *U* and *V* colour difference signals at the

demodulator outputs. These can then be matrixed to produce RGB according to:

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} 1 & 0 & \dfrac{1}{0.877} \\ 1 & \dfrac{-0.114}{0.587 \cdot 0.493} & \dfrac{-0.299}{0.587 \cdot 0.877} \\ 1 & \dfrac{1}{0.493} & 0 \end{bmatrix} \cdot \begin{bmatrix} Y \\ U \\ V \end{bmatrix}$$

The circuit arrangement needed to produce these relationships is shown in Fig. 36. Alternatively, the digital component signals, Y, C_R and C_B can be produced by scaling U and V according to:

$$C_B = \frac{224}{219} \cdot \frac{0.5}{0.886} \cdot \frac{1}{0.493} U$$

and

$$C_R = \frac{224}{219} \cdot \frac{0.5}{0.701} \cdot \frac{1}{0.877} V$$

with the factor 224/219 arising because the Y coding range is slightly less than that used for C_R and C_B.

However, C_R and C_B can be produced directly by including these gain factors in the demodulator using scaled sine and cosine waveforms (Section 2.2). In this case, the matrix to produce RGB becomes:

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} 1 & 0 & \dfrac{219}{224} \cdot \dfrac{0.701}{0.5} \\ 1 & \left(\dfrac{-219}{224} \cdot \dfrac{0.886}{0.5} \cdot \dfrac{0.114}{0.587}\right) & \left(\dfrac{-219}{224} \cdot \dfrac{0.701}{0.5} \cdot \dfrac{0.299}{0.587}\right) \\ 1 & \dfrac{219}{224} \cdot \dfrac{0.886}{0.5} & 0 \end{bmatrix} \cdot \begin{bmatrix} Y \\ C_B \\ C_R \end{bmatrix}$$

Another alternative is to simplify the matrix by using scaling factors of 1/0.493 and 1/0.877 in the demodulators to produce $B - Y$ and $R - Y$ signals directly. The relationship for RGB then becomes:

$$\begin{bmatrix} R \\ G \\ B \end{bmatrix} = \begin{bmatrix} 1 & 0 & 1 \\ 1 & \dfrac{-0.114}{0.587} & \dfrac{-0.299}{0.587} \\ 1 & 1 & 0 \end{bmatrix} \cdot \begin{bmatrix} Y \\ B - Y \\ R - Y \end{bmatrix}$$

which requires only two non-unity coefficients. Scaling is then required in the digital component outputs according to:

$$C_B = \frac{224}{219} \cdot \frac{0.5}{0.886} \cdot (B - Y)$$

and

$$C_R = \frac{224}{219} \cdot \frac{0.5}{0.701} \cdot (R - Y)$$

Additional features are sometimes required to take account of the coding ranges used in the a–d and d–a converters. Ideally, a nine-bit a–d converter would be used with black and white still corresponding to levels 16 and 235, but now within a coding range from −128 to +383. The standard[1] eight-bit coding ranges for Y, R, G and B would then be produced directly, with the additional bit providing range for the positive and negative excursions of chrominance and syncs. At present, however, the high cost of a nine-bit converter makes its use unattractive. Because of this, it is usual to use an eight-bit converter, as mentioned in Section 4.2, with black stabilised at level 64 and white set at level 204. The composite signal then extends from level 4 at the sync. bottom to level 251 at the most positive chrominance excursion. This almost minimises the quantising distortion, but leaves virtually no headroom for highly coloured overloads. With the eight-bit converter, an additional scaling factor of 219/140 is required in all three channels. While this can be combined with other factors in the colour difference channels, this requires an additional code conversion process in the luminance channel to provide the correct gain and black level.

Because of the large gain factors involved, cross-colour signals produced by the demodulators can go well beyond the normal range of RGB signals, particularly in the B channel. Some reduction of the worst cross-colour can therefore be produced by limiting the signals at the matrix output to the normal R, G and B ranges. While this is generally beneficial, it should be noted that this reduces the reversibility of the decoding process and, if carried out in the sampled domain, leads to aliasing of the harmonic components produced by limiting. The latter effect can be avoided by using extra headroom in the d.a.c.'s and limiting in the analogue domain.

Synchronising information required for the generation and insertion of timing reference signals can be obtained from the divider circuit of the line-locked sampling loop described in Section 4.2.2. Analogue mixed synchronising pulses to accompany

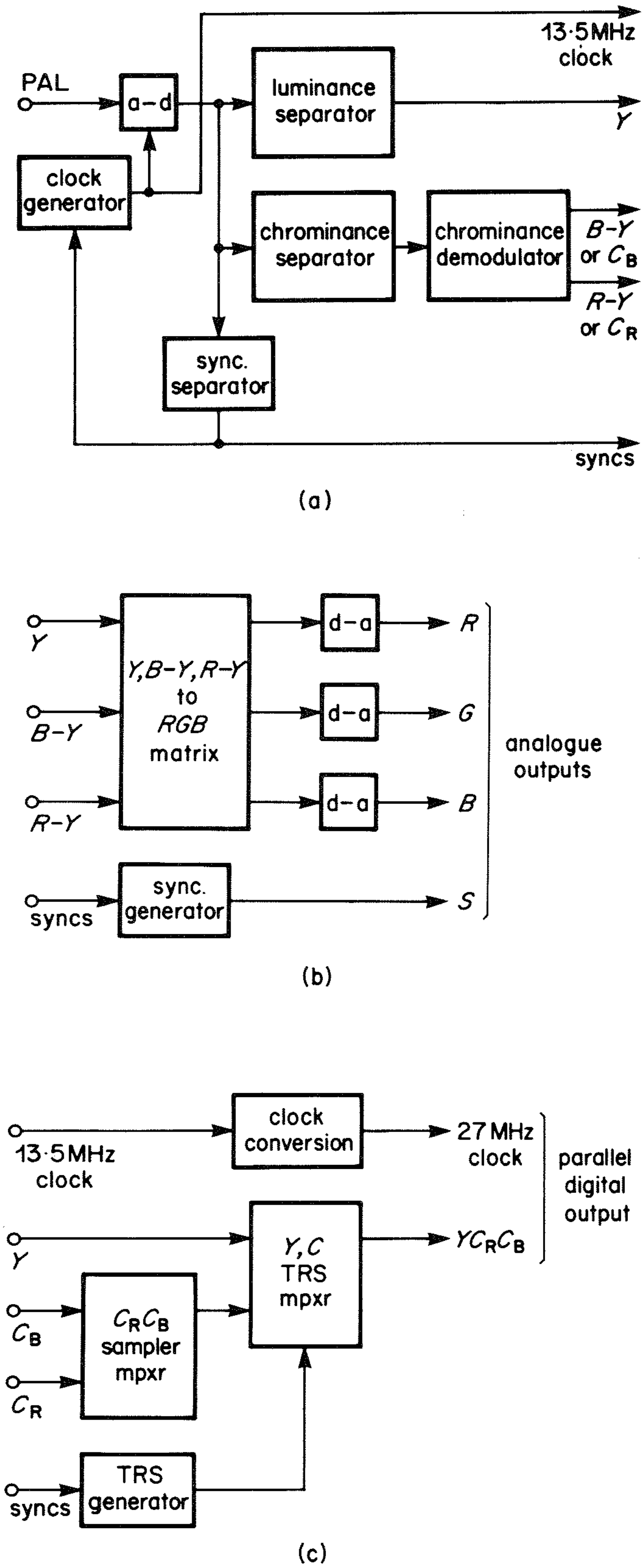


Fig. 35 – Alternative output configurations for digital colour decoding: (a) common circuitry, (b) an analogue RGB output interface and (c) a parallel digital output interface.

the *RGB* outputs can be regenerated using the method described in Section 3.2.3.

4.4. Modulated chrominance processing techniques

The PAL demodulation process is inherently phase-sensitive. Because of this, some post-demodulation chrominance-luminance separation methods that can be used in the arrangement of Fig. 23(c) cannot be reproduced using conventional filtering techniques before the demodulators in the simpler arrangement of Fig. 23(b). Instead PAL modifiers and phase-shift filters have to be included to alter the chrominance contributions from different lines so that all have the same reference phase.

4.4.1. PAL modifiers

The alteration of subcarrier phase and *V*-switch sense is obtained using a PAL modifier. This consists of a multiplier fed with a twice subcarrier frequency sine wave, followed by a chrominance band-pass filter, as shown in Fig. 37. The phase of the twice subcarrier frequency sine wave relative to the reference phase of the PAL signal determines the phase

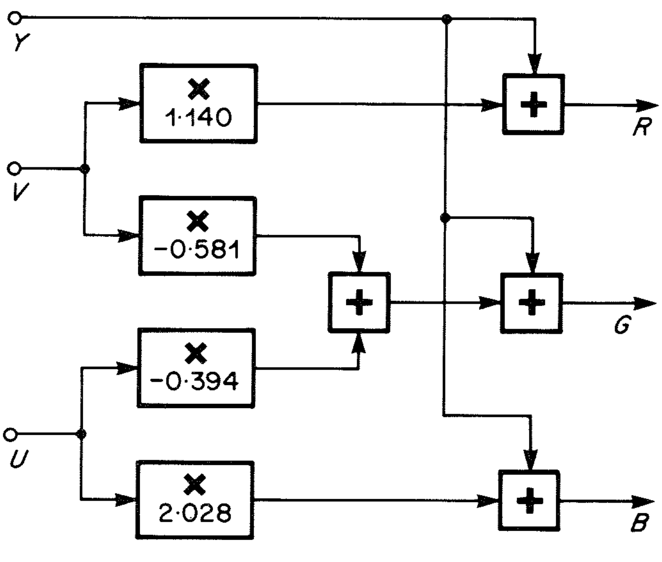


Fig. 36 – A YUV to RGB matrix.

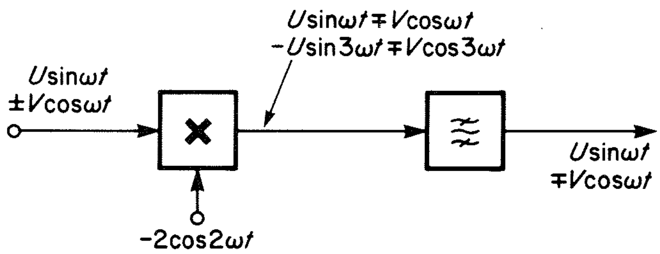


Fig. 37 – A PAL modifier.

shift of the modified signal. As drawn, with a PAL chrominance signal represented by $U \sin \omega t \pm V \cos \omega t$ and multiplied by $-2 \cos 2\omega t$, there is no shift of the reference phase, but the V-switch sense is inverted.

Alternatively, if $2 \sin 2\omega t$ is fed to the multiplier, then a 90° phase shift is produced as well as the V-switch inversion, since:

$$
\begin{aligned}
&(U \sin \omega t \pm V \cos \omega t)(2 \sin 2\omega t) \\
&= U \cos \omega t \pm V \sin \omega t \\
&\quad - U \cos 3\omega t \pm V \sin 3\omega t \\
&= U \sin(\omega t + 90^\circ) \mp V \cos(\omega t + 90^\circ) \\
&\quad - U \cos 3\omega t \pm V \sin 3\omega t
\end{aligned}
$$

and the terms at three-times subcarrier frequency are removed by the chrominance band-pass filter.

The action of a modifier is shown in spectral terms in Fig. 38. Fig. 38(a) represents the baseband spectrum of an analogue video signal with chrominance in the shaded region around the colour subcarrier frequency and Fig. 38(b) represents the spectrum of the twice subcarrier frequency sine wave used in the modifier. Convolving these spectra produces the result shown in Fig. 38(c) in which the chrominance information from f_{sc} is shifted to $-f_{sc}$ and that from $-f_{sc}$ is shifted to f_{sc}.

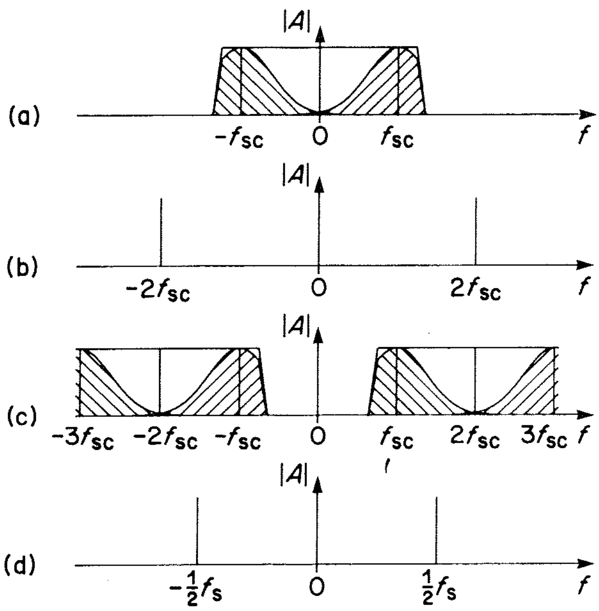


Fig. 38 – Frequency spectra for PAL modification: (a) the spectrum of a composite video signal, (b) the spectrum of the twice subcarrier frequency sine wave used in the modifier, and (c) the result of convolving (a) and (b). The half sampling frequency positions for 13.5 MHz sampling are shown in (d) relative to the modified components in (c).

If this same operation were performed in a sampled system, the resulting spectrum would be that of Fig. 38(c) folded back for frequencies above half the sampling frequency. The positions of the half sampling frequency are shown relative to the modified spectrum in Fig. 38(d). Although the required chrominance modification would have been achieved, aliased low frequency luminance signals centred on $2f_{sc}$ in Fig. 38(c) would occupy the same band as chrominance at f_{sc}. For sampling frequencies at or near to $3f_{sc}$, for example 13.5 MHz, this can be avoided by inserting a chrominance band-pass filter before the multiplier. Fig 39 shows the repeated spectra of a modifier system, including prefiltering, for a sampling frequency slightly greater than $3f_{sc}$. Fig. 39(a) shows the spectrum of the sampled video signal, whilst in Fig. 39(b) the low frequency luminance has been removed by a chrominance band-pass filter. When the spectrum of Fig. 39(b) is convolved with the sampled twice subcarrier frequency sine wave, Fig. 39(c), the modified spec-

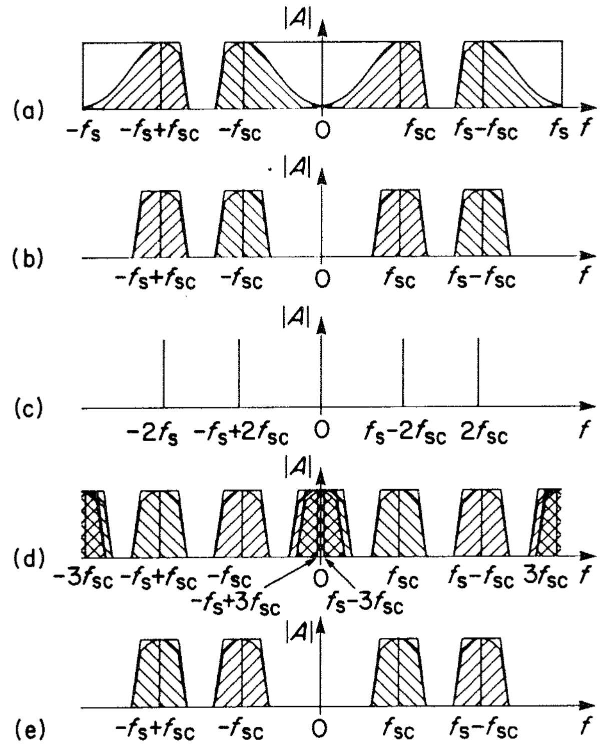


Fig. 39 – Frequency spectra for a PAL modifier operating at a sampling frequency of 13.5 MHz: (a) the spectrum of sampled video signals, (b) sampled chrominance signals obtained by band-pass filtering, (c) the spectrum of a sampled twice subcarrier frequency sine wave, (d) the result of convolving (b) and (c), and (e) the spectrum of PAL-modified chrominance obtained by band-pass filtering (d).

trum of Fig. 39(d) is produced. Finally, the unwanted products can be removed by a second chrominance band-pass filter to leave the modified spectrum as shown in Fig. 39(e).

The implementation of a PAL modifier in digital circuitry uses the same four-quadrant multiplier techniques used in the demodulator, except that the read-only memory contains an appropriately phased sine wave of twice the frequency. Although the subcarrier phase information is again derived from the demodulator reference phase, a constant phase offset is added to take account of processing delays between the modifier and the demodulators.

By considering the processes involved it can be shown that the combination of a demodulator and a remodulator is sometimes equivalent to a PAL modifier. If the colour difference signals are remodulated in the same phase as they were demodulated, the circuit has the effect of a band-pass filter. However, if the V signal is inverted at some stage in the demodulation and remodulation process, then the effect is that of a PAL modifier. These relationships, shown in Fig. 40, allow an equivalent filtering action to be produced either before or after demodulation.

A considerable disadvantage in the use of PAL modifiers is that their operation is adversely affected by differential phase distortion. This is because the phase shift introduced by a modifier is a function of the subcarrier reference phase in the demodulator. As differential phase distortion alters the relationship between the reference burst and the active-line chrominance signals, modified signals no longer match the phase of those from unmodified lines. In luminance separation comb filters, the cancellation of the subcarrier components depends on the phases of the individual contributions from different lines being accurately matched. Consequently, if a modifier is used, phase distortion leads to incomplete subcarrier suppression, even in plain areas. This is particularly serious for decoding applications where subsequently the signals are to be recoded to PAL, such as at the inputs to a digital component studio. Then the residual subcarrier interferes with the subcarrier in the following coder to cause a hue error when the signal is finally decoded.

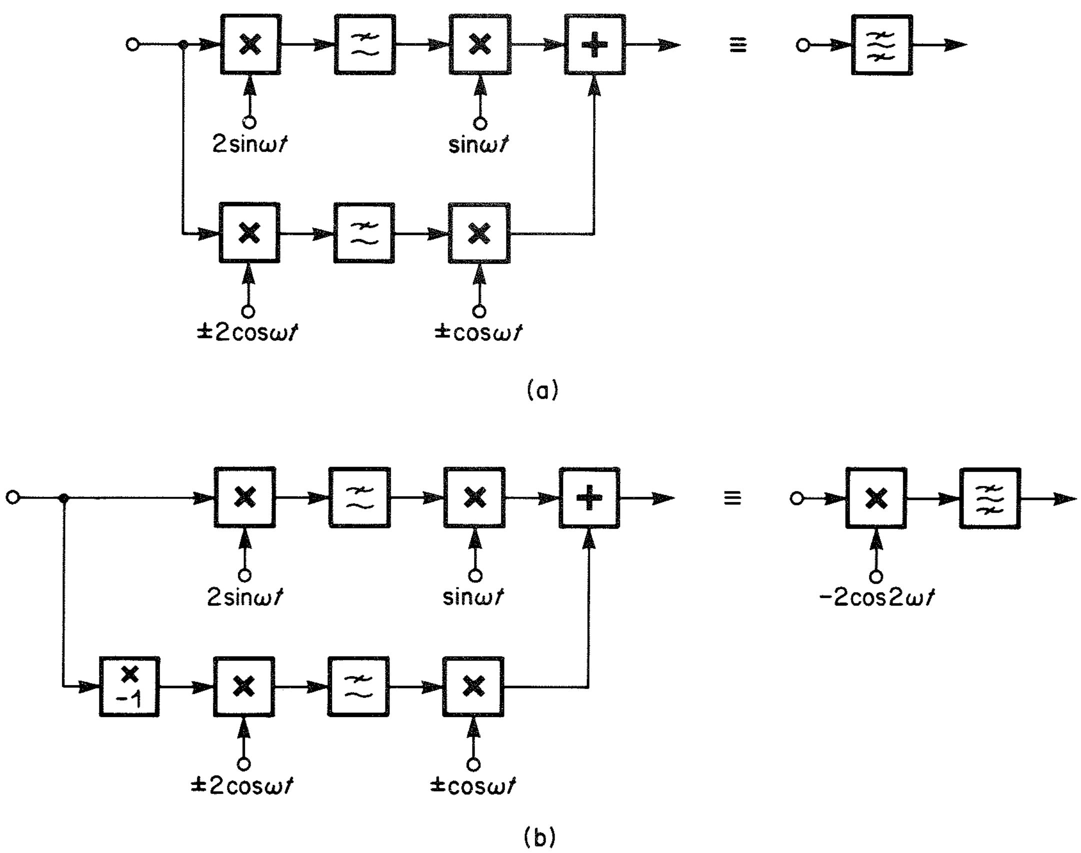


Fig. 40 – The effect of decoding and recoding: (a) with modulation in the same phase as demodulation and (b) with the V signal remodulated with the opposite phase to demodulation.

4.4.2. Phase-shift filters

While the effect of differential phase distortion makes modifiers unattractive for luminance separation, there is no deleterious effect for chrominance. On the contrary, combining oppositely switched lines together to produce the chrominance signal for demodulation ensures suppression of Hanover bars. However, when oppositely switched lines are only used for obtaining chrominance, the complication of a modifier is unnecessary because an equivalent effect can be produced more simply. By using separate inputs to the U and V demodulators, the V-phase contributions can be inverted, where necessary, without affecting the U-phase signals. Also the phase shift of 90° required in some contributions for both U and V signals can be obtained by a special form of band-pass transversal filter.

The symmetrical transversal filters described in Section 3.2.4. produce no phase shift of one frequency relative to another. However, if a filter with anti-symmetrical taps and no contribution from the centre term is used, all frequencies are shifted by 90°. As the amplitude characteristic is a summation of harmonic sine wave components weighted by the filter coefficients, the response at zero frequency is necessarily zero. Thus, only band-pass or high-pass characteristics are possible. As in Section 3.2.4, appropriate coefficient values for the desired response can be obtained by solving a set of simultaneous equations,[29] except that in this case each equation is a summation of sine terms instead of cosines.

A digital filter with anti-symmetrical taps can be implemented using a similar ladder structure to that for a filter with symmetrical taps shown in Fig. 18. Anti-symmetry is obtained by altering the centre-term stage to give zero contribution and to invert the contributions added in the upper delay line. This arrangement is shown in Fig. 41.

4.5. NTSC demodulation

Because of the similarity between the two systems, many of the features described for PAL demodulation can also be used for the NTSC system. Several of the relevant differences in NTSC signals have been included in Section 3.3 on NTSC encoding. However, there are some additional features which are specific to NTSC demodulation.

The NTSC chrominance signal, represented by the expression:

$$Q \sin \omega t + I \cos \omega t$$

can be demodulated by multiplying by $2 \sin \omega t$ and $2 \cos \omega t$ and low-pass filtering the resulting products. As Q is a low bandwidth (0.5 MHz) double sideband signal and I is a wideband (1.3 MHz) asymmetrical sideband signal, different Q and I low-pass filters are required. For Q, a filter similar to that used in the coder can be used (Fig. 21(b)) with the sharp cut-off ensuring that crosstalk from high frequency I signals is removed. For I, a slow roll-off filter with a somewhat narrower bandwidth than the wideband I filter used in the coder is preferred. Some gain in this filter in the range 0.5 to 1.3 MHz could be included to compensate partially for the sideband asymmetry of the I signal. However, this is of doubtful benefit because of the increase in cross-colour that would result.

After demodulation, the axis rotation of the I and Q signals has to be removed using the relationships:

$$U = Q \cos 33° - I \sin 33°$$

and

$$V = Q \sin 33° + I \cos 33°$$

These relationships could be implemented with the matrix arrangement of Fig. 42(a) or for RGB out-

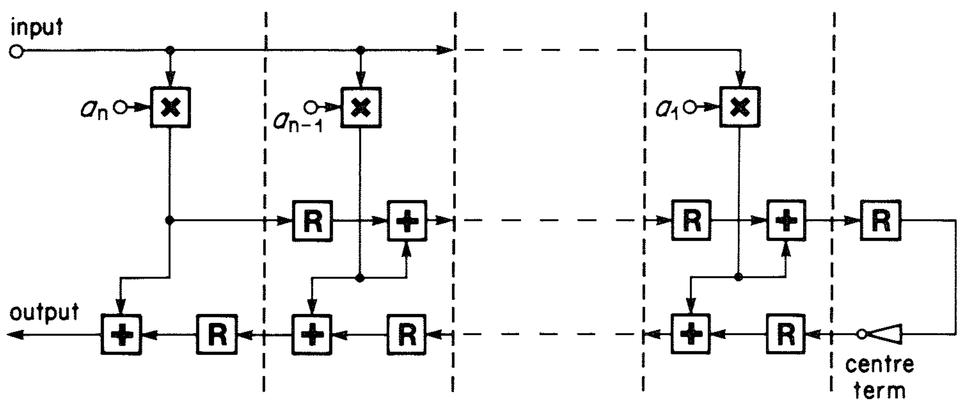


Fig. 41 – Phase shift filtering: using the alternative centre term stage shown here produces a characteristic which has a 90° phase shift at all frequencies with the same modular filter structure as in Fig. 18.

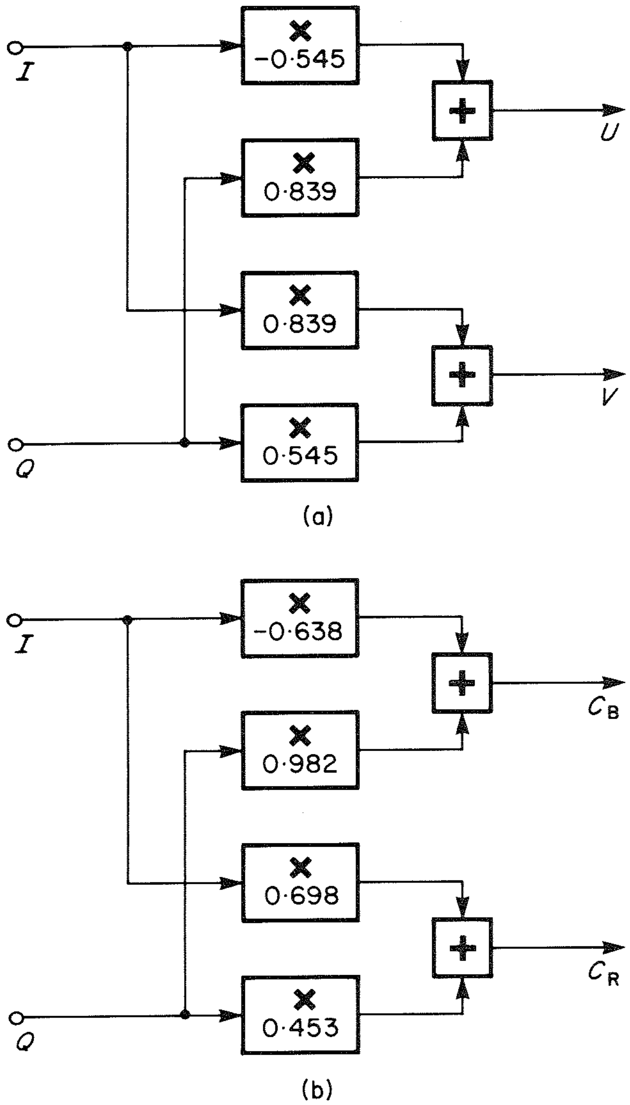


Fig. 42 – Matrix circuits for producing (a) U and V or (b) C_B and C_R colour difference signals from I and Q.

puts, could be incorporated in the *YUV* to *RGB* matrix of Fig. 36. Alternatively, for digital component signals, the additional relationship between *U*, *V* and C_B, C_R signals (Section 4.3.3) could be incorporated into the matrix coefficients producing the values shown in Fig. 42(b).

In the digital phase-locked loop only minor modifications are required to change from PAL to NTSC operation. As the NTSC signal includes no *V*-switch modulation, the PAL switch bistable must be disabled. The loop can then lock directly to the reference phase (the $+U$ axis) using as the error signal the amount of the $-U$ burst signal demodulated in the *V* channel. The 33° phase rotation required to generate *I* and *Q* subcarriers is then provided by a constant offset added to the reference phase signal during the active-line period. It should be noted that, although the counting ratios described in Section 2.1.2 are relatively simple for NTSC signals, the full accuracy of the 16-bit lower stage counter is required in the demodulator. This is necessary in order to provide sufficient precision in the frequency control of the digital loop.

Similar 'broad-band' techniques to those described for NTSC coders in Section 3.3 can also provide considerable simplifications in an NTSC demodulator. As in the coder, ignoring the sideband asymmetry allows the multipliers to demodulate *U* and *V* directly and to use identical low-pass filters, thus avoiding the need for an *I*, *Q* matrix and the 33° phase offset from the burst. This approach provides much greater similarity to the PAL demodulator.

5. Conclusions

The use of line-locked sampling at 13.5 MHz in digital composite colour coding and decoding equipment has the advantage of providing direct compatibility with digital studio component signals. Although somewhat more complicated than subcarrier-locked methods, line-locked techniques can accept non-mathematical and monochrome signals and do not require a subsequent sample rate conversion process. The digital methods have the advantages over conventional analogue techniques of more stable and better-defined performance.

The methods of coding and decoding described both involve the generation of quadrature subcarrier signals from the line-locked sampling frequency by a system of ratio counters. The reference phase signal produced by the counters is used to address sine and cosine read-only memories which generate the subcarrier waveforms. Suitable accuracy can be obtained with an eleven-bit reference phase signal. For coders, modulation is achieved using four-quadrant digital multipliers.

In decoders using field or picture delay comb filters, the generation of stable line-locked sampling pulses is important. Improved stability can be obtained using a digital phase detector to lock the sampling rate to the line pulses.

The line-locked digital demodulator employs ratio counting and multiplier techniques similar to those used for the coder with the additional feature of stabilisation of the local reference phase to the incoming colour bursts. Subcarrier frequency control for phase locking is obtained by altering the counting ratios by a small amount in response to the phase error. Non-mathematical operation is improved by including an integrator in the frequency control loop filter to avoid the need for a non-zero phase error.

The modulation and demodulation techniques described can be combined with luminance and chrominance filters in a wide range of configurations for coding or decoding. Similar circuitry can be used for both PAL and NTSC, particularly if the side-band asymmetry of chrominance in the NTSC system is ignored. Several prototype line-locked PAL and NTSC decoders using these techniques have been produced for investigations into chrominance-luminance separation methods and the technique has been developed for use in the Television Service by Designs Department.

6. Acknowledgement

The author wishes to acknowledge the contribution made by P.W. Fraser to the development of digital hardware for coding and decoding in support of the techniques described in this report.

7. References

1. Encoding parameters of digital television for studios. CCIR Recommendation 601, XVth Plenary Assembly, Geneva 1982, **Vol. XI,** Part I, pages 271–273.

2. DEVEREUX, V.G. 1970. Pulse code modulation of video signals: 8-bit coder and decoder. BBC Research Department Report BBC RD1970/25.

3. DEVEREUX, V.G., 1971. Pulse code modulation of video signals: subjective study of coding parameters. BBC Research Department Report BBC RD1971/40.

4. WALKER, R., 1971. Digital line-store standards conversion: a feasibility study. BBC Research Department Report BBC RD1971/44.

5. DEVEREUX, V.G., 1973. Digital video: differential coding of PAL signals based on differences between samples one subcarrier period apart. BBC Research Department Report BBC RD1973/7.

6. PHILLIPS, G.J. *and* CHAMBERS, J.P., 1975. Reduction of Errors in Television Systems: Improvements in Digital Transmission of Television. UK Patent Specification GB 1 415 519. 26th September 1975.

7. CHAMBERS, J.P., 1974. The use of digital techniques in television waveform generation. IEE Conference Publication No. 119 (IBC 74), pp. 40–46.

8. DEVEREUX, V.G., 1975. Digital video: differential coding of PAL colour signals using same line and two-dimensional prediction. BBC Research Department Report BBC RD1975/20.

9. DEVEREUX, V.G., 1975. Digital video: sub-Nyquist sampling of PAL colour signals. BBC Research Department Report BBC RD1975/4.

10. WESTON, M., 1976. A PAL/YUV digital system for 625-line international connections. BBC Research Department Report BBC RD1976/24.

11. REID, D.F., STOTT, J.H. *and* OSBORNE, D.W., 1977. Experimental digital transmission of multiplexed video and audio signals at 60 Mbit/s through a satellite. BBC Research Department Report BBC RD1977/32.

12. OLIPHANT, A., 1980. Weston Clean PAL. BBC Research Department Report BBC RD1980/1.

13. BALDWIN, J.L.E. *and* LEVER, I.R., 1978. Proposed digital television standards for 625 line PAL signals. IEE Conference Publication No. 166 (IBC 78), pp. 214–217.

14. DREWERY, J.O., 1984. The present status of Clean PAL systems. BBC Research Department Report BBC RD1984/6.

15. BALDWIN, J.L.E., 1976. Sampling frequencies for digital coding of television signals. IBA Technical Review, **Vol. 9,** pp. 32–35.

16. CLARKE, C.K.P., 1982. Digital standards conversion: comparison of colour decoding methods. BBC Research Department Report BBC RD1982/6.

17. ASTLE, J.M. *and* SHELTON, W.T., 1980. Digital television standards converter with random access memory storing four fields. IEE Conference Publication No. 191 (IBC 80), pp. 159–163.

18. CLARKE, C.K.P., 1983. Apparatus for Digital Demodulation or Modulation of Television Chrominance Signals. UK Patent Specification GB 2 059 711. 20th July 1983.

19. CLARKE, C.K.P., 1983. Apparatus for Digital Demodulation or Modulation of Television Chrominance Signals. UK Patent Specification GB 2 061 053. 10th August 1983.

20. JONES, A.H., 1980. Digital television standards. IEE Conference Publication No. 191 (IBC 80), pp. 79–82.

21. Characteristics of television signals. CCIR Report 624–2, XVth Plenary Assembly, Geneva 1982, **Vol. XI,** part I.

22. CHAMBERS, J.P., 1976. The Relation of Frequencies in Integer Ratio. UK Patent Specification GB 1 455 821. 17th September 1976.

23. DEVEREUX, V.G., 1984. Filtering of the colour-difference signals in 4:2:2 YUV digital video coding systems. BBC Research Department Report BBC RD1984/4.

24. Bit-parallel digital interface for component video signals in 525-line and 625-line television systems. Annex to CCIR Draft Report AG-11.

25. Specification of television standards for 625-line System I transmissions, 1971. Jointly published by BBC, ITA and BREMA.

26. DREWERY, J.O., 1983. Colour Television Transcoding Matrices. UK Patent Specification GB 2 073 535.

27. WESTON, M., 1980. A set of time-varying television test patterns. BBC Research Department Report BBC RD1980/9.

28. MACDIARMID, I.F., 1959. Waveform distortion in television links. *The Post Office Electrical Engineer's Journal*, **Vol. 52,** July 1959, pp. 108–114.

29. CLARKE, C.K.P., 1977. Digital television: the use of waveform estimates in error correction. BBC Research Department Report BBC RD1977/27.

30. LENT, S.J. *and* REED, C.R.G., 1974. Reduction of cross-colour effects in PAL (System I) colour television by the use of low-pass and notch filters: subjective tests. BBC Research Department Report BBC RD1974/25.

31. CLARKE, C.K.P., 1982. High quality decoding for PAL inputs to digital YUV studios. IEE Conference Publication No. 220 (IBC 82), pp. 363–366.

32. DORWARD, R.M., 1969. Aspect of the quantisation noise associated with the digital coding of colour television signals. *Electronics Letters*, **Vol. 6,** No. 1, January 1970, pp. 5–7.

33. DEVEREUX, V.G., 1982. Tests on eight video p.c.m. codecs in tandem handling composite PAL and monochrome video signals. BBC Research Department Report BBC RD1982/19.

34. GARDNER, F.M., 1967. *Phase Lock techniques*. Wiley, New York, 2nd Edition, 1967.

Printed by BBC RESEARCH DEPARTMENT, Kingswood Warren, Tadworth, Surrey, KT20 6NP